INSTRUCTION MANUAL FOR THE BA-5 ATTENUATION CALIBRATOR

Weinschel Engineering

KENSINGTON, MARYLAND

Weinschel Engineering Co., Inc.

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INSTRUCTION MANUAL FOR THE

BA-5 ATTENUATION CALIBRATOR

By
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Keyboard Oscillator
Attenuation Standards

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GENERAL INTRODUCTION

1.1 Scope of Manual.

This manual covers the description, theory of operation, operating procedure, and maintenance instructions for BA-5 Bolometer Amplifier and Attenuation Calibrator.

1.2 Purpose of Equipment.

The BA-5 Attenuation Calibrator is a combination of a constant current D.C. bias supply for a barretter, a precision audio attenuator, a high-gain 1,000 cps bolometer amplifier and an indicator. The equipment was designed for general use in measuring r-f power level ratios in the frequency range from 20 mc to 40,000 mc or at other frequencies where suitable bolometer mounts are available for use with the coaxial or waveguide transmission lines used at the frequency of measurement.

It is particularly suited to the measurement of r-f power level ratios in applications such as antenna pattern measurements, insertion loss measurements and R.F. amplifier gain measurements, when used in conjunction with other equipment described in Table 1 on Page 23 and in the block diagrams shown on the attached drawings B-114-103, B-114-121 and B-114-169. Because of the low noise level of the amplifier, these measurements can be performed at low power levels in the order of 200 microwatts peak power 100% square wave modulated at 1,000 cps which are easily obtained from R.F. signal sources without the risk of burning out bolometer elements. It is possible to measure directly r-f power ratios of 30 db without changing amplifier gain or use of R.F. substitution attenuators.

The audio attenuator is placed ahead of the amplifier-indicator combination. This permits operation of the amplifier at a constant input level, thereby eliminating errors due to possible non-linearity of amplifier and inaccuracy of the indicator system.

This makes it possible to make measurements approaching the accuracy of the precision audio attenuator. The inaccuracy of the audio attenuator, which is due to the accuracy of the resistance values at DC (.1%) and the response at 1000 cps (finite phase angle of the resistors and crosstalk) does not exceed .02 db, which corresponds to an inaccuracy of the R.F. measurement of .01 db R.F. The readability of the indicator and the random amplitude changes of the R.F. source in a single channel measurement (Dwg. B-114-103) usually do not permit the making of measurements to such high accuracies. When using relatively stable sources such as our MS-1 (50-250 mc), MS-2 (250-920 mc) or MS-3 (900-2000 mc), losses to 20 db can be measured readily to an accuracy of .05 db. Due to the influence of noise, the accuracy for measurements of R.F. losses to 30 db is about .1 db.

In order to achieve the high accuracies inherent in and possible with this design, a more sensitive indicator and an automatic compensation for R.F. level changes is required. Dwg. B-114-121 illustrates a two-channel method where the R.F. is split into a measurement channel and a reference channel for "bucking." Even if the amplitude of the R.F. source varies during the measurement by plus or minus .2 db, insertion loss measurements between 0-10 db are readily repeated within .01 db and measurements to 20 db are repeated within .01 and .02 db.

Because the insertion loss depends upon load and source impedance, both must be well defined in order to make precision insertion loss measurements. This is possible in 3/8" coaxial systems to 2000 mc, 7/8" coaxial systems to 4000 mc and between 2000 and 12,000 mc in common waveguide transmission lines. Then accuracies of .02 db/10 db can be realized.

Because the precision A.F. attenuator is between the bolometer mount and the amplifier input, a relatively high A.F. voltage originating in the barretter can be attenuated so that it will not overload the amplifier.

If the error due to deviation from square law of a barretter is to be kept to about .01 db, a maximum peak power of 200 microwatts R.F. (100% square wave modulated at 1,000 cps) may be used in a FRD 610 barretter. Due to the influence of noise on linear detectors and the noise arising in the bolometer amplifier, an error of about .1 db is encountered when measuring a 30 db R.F. loss. It is permissible to apply safely 1 milliwatt R.F. peak power to the FRD 610 barretter. This increases the measurement range by 7 db while causing a deviation from square law of about .07 db. (Add .07 db to measured attenuation for correction).

Using, e.g., a Sperry or Narda 821 barretter, it is possible to apply 6 db more R.F. before encountering the same deviation from square law because the D.C. bias power is increased 6 db. However, due to the difference in the thermal time constants, the demodulation efficiency at 1,000 cps is 4 db (A.F.) lower corresponding to a 2 db R.F. drop. By using 800 microwatt R.F. peak power, the range of measurement has been extended to about 34 db with a maximum error of .01 db R.F. due to deviation of the barretter from square law.

1.3 Specifications.

1.3.1 Dimensions: Standard relay rack mounting, black crackle panel, 10½" x 19". Maximum depth behind panel 13½". Weight 35 lbs.

1.3.2 Ambient temperature: 25°C plus or minus 10°C.

1.3.3 Power Requirements: 115 volts plus or minus 10%, 60 cps only, 190 VA.

1.3.4 Fuse: 3 amp. Bussman AGC, mounted in holder on rear of chassis.

1.3.5 Connectors: Panel Mounted: Input from "Bolometer" BNC female, panel jack; "Output" to external meter, BNC female, panel jack; Chassis Mounted: Power line input, standard 115 volt, 2 wire recessed male cap on rear of chassis.

1.3.6 Controls: Panel Mounted: An "On-Off" switch controls AC power, ruby pilot light indicates power "on." Three 2" skirt knobs marked "R.F. DB" cover the range 0-10 db in ten 1 db steps; 0-2 db in one 1 db step from 0-1 db and ten .1 db steps from 1-2 db; minus .01 db to plus .11 db continuous, marked in .01 db steps. 4 toggle switches of 10 db each. Two 1" skirt knobs serve as gain controls. The "coarse" control has 7 steps of about 10 db (A.F.) each; the "fine" control covers continuously a range of about 10 db (A.F.). A toggle switch changes the "meter time-constant" from "short" to "long". Screwdriver control marked "adjust" for setting the bolometer bias current between 0 and 10 MADC.

Chassis Mounted: On top of chassis (P-1), B_+ adjustment; below chassis, attenuator termination control (P-5); both are pre-set, must not be changed.

1.3.7 Voltage Gain: Maximum 50×10^6 adjustable over range of 80 db A.F. gain changes less than plus or minus .1 db through the range of line voltage from 105-130 volts.

1.3.8 Audio Frequency: 40 cps bandpass centered around 1000 cps.

1.3.9 Maximum Output Voltage: For F.S. meter deflection, 1.5 v RMS through a cathode follower is available at "output" terminal.

1.3.10 Input Impedance: About 2000 ohms for use with bolometers having a D.C. resistance between 50 and 200 ohms.

1.3.11 Output Impedance: About 250 ohms, recommended load impedance for minimum distortion above 10,000 ohms.

1.3.12 Equivalent Noise: About 30 millimicrovolts RMS referred to the input when connected to a shielded 200 ohm wirewound resistor.

1.3.13 Bolometer Bias Current: 0-10 MADC, adjustable, constant current type.

1.3.14 Protection of Bolometer Against Burnout by Transients of Bias Current: Connection or disconnection of bolometer while the preamplifier is "on" will not damage bolometer. Switching power "off" or "on" while bolometer is connected to preamplifier will not damage bolometer.

1.4 Description.

The Model BA-5 Bolometer Amplifier and Attenuation Calibrator consists of a 1000 cps, low noise, bolometer amplifier, combined with a precision A.F. attenuator and a vacuum tube voltmeter with VSWR and db scales. A cathode follower output for external meters permitting the use of the BA-5 in R.F. two channel circuits in combination with the ND-1 Differential Null Detector is provided. It permits the measurement of R.F. power level ratio to 30 db using recommended accessories shown in block diagrams B-114-103, B-114-121, B-114-169 and Table 1, page 23.

The amplifier has a self-contained bias supply for the bolometer, together with a meter for monitoring the bolometer current. The circuit is designed to enable the bolometer element to be connected or disconnected without turning off the bias supply and without danger of burning out the bolometer element.

The bolometer current supply is of the "constant current" type. Even a considerable change in the bolometer resistance due to R.F. overload will not cause any noticeable change of the bolometer current.

If the R.F. power is 100% square wave modulated at 1,000 cps, there is a simple way to prevent an R.F. overload. With the bias current set to the specified value, adjust both "coarse" and "fine" gain to minimum and all BA-5 attenuators to zero. Increase R.F. power gradually until meter reads full scale. This level corresponding to an input of about 2.5 millivolts RMS at 1,000 cps assures one that there is no overload with most presently available bolometers supplied by F. & R. Machine Works, Narda, P.R.D. and Sperry.

The bias control covers a range of 0-10 MADC. It can be turned completely off if it is desired to use a crystal probe.

The amplifier has a low noise level which makes it possible to obtain measurements at a low r-f power level, which is usually available and further protects the bolometer element against accidental burnout. This low level of noise has been obtained by careful design of the input circuits; the amplifier can be operated in the vicinity of a-c power supplies and other 60 cycle equipment without danger of excessive hum pickup by the amplifier.

The input resistance of the amplifier is about 2000 ohms which is relatively high compared with the bolometer element resistance to insure a negligible change of the voltage applied to the amplifier through variations of bolometer resistance caused by r-f power level changes.

The voltage from the bolometer is measured by means of a calibrated attenuator before amplification. This means that amplifier linearity does not enter into the accuracy of the results obtained.

A meter time constant switch provides a choice of "short" or "long" to take care of conditions of noise, or other interference which causes erratic meter movement.

The frequency response of the amplifier is determined by the twin-tee networks resulting in a nearly flat bandpass centered around 1,000 cps.

1.5 Main Uses.

The BA-5 is well adapted for R.F. attenuation or antenna pattern measurements or as calibrated indicator for impedance measurements in combination with slotted lines using either a barretter or a crystal diode. However, if a crystal detector is used, the dc bias supply to the input must be switched off. This is done by rotating the screwdriver-control potentiometer accessible through a hole in the front panel to its extreme counterclockwise position. If this precaution is not taken, the crystal might be damaged by the dc bias current. When using a crystal diode, it is essential that the crystal be operated in the square law region of its response characteristic.

1.6 Unpacking and Preparation For Use.

The BA-5 is shipped in a double carton with three (3") inches of packing material between the inner and outer cases. Normal care should be exercised in removing the instrument from the case and removing packing material. There should be no loose items with the exception of the line cord and the Instruction Manual.

Tubes are shipped in place.

After unpacking the instrument, examine it for mechanical damage. If any is found, notify the shipping company immediately.

1.6.1 Remove the bottom cover of the chassis and examine the leads connecting to the two floating chassis, to insure that none of these leads has been broken in transit. Be particularly sure to check the ground connections since erratic grounds can produce severe oscillation conditions which render the instrument useless.

Check the condition of the 3 amp Bussman AGC fuse. Be sure that all tubes are firmly seated. Remove the damping wire which is wrapped around and across the terminals of the 200 microamp output meter.

Turn the bolometer bias control to its maximum counter-clockwise position.

1.6.2 The other equipment required will depend on the purpose for which the instrument is to be used. (See Section 3). Although the instrument is thoroughly shielded, endeavor to keep it away from high intensity power frequency or r-f fields. To insure greatest accuracy, see that the line voltage is within the operating limits. If necessary, use an A.C. voltage regulator to provide the proper AC line voltage.

1.6.3 The main precaution to be observed when using this equipment is to keep the coaxial input cable from the bolometer to the input of the BA-5 as short as conveniently possible. Under no circumstances, allow it to exceed three feet in length. Do not permit any coaxial cable to pass through a strong magnetic field at the power or modulating frequency since the shield of the cable is ineffective at such low frequencies for magnetic disturbances.

Do not place the BA-5 next to the audio modulator for the R.F. source. Avoid ground loops when connecting your shielded cables. Do not place the BA-5 in the same metal rack with the audio modulator. When laying out the equipment, take care to keep the cable well away from transformer fields and other high intensity magnetic areas. If a stabilized voltage regulator is used, keep it as far away as conveniently possible from the test set-up.

1.6.4 Set the "coarse" and "fine" gain controls to their maximum counterclockwise position. Plug in the line cord and switch on the power. Allow the equipment to warm up for five minutes. Plug the bolometer cable into "Input" and set the "Bolometer Current" control by means of a screwdriver through the hole in the front panel marked "Adjust" until the "Bolometer Current" meter indicates the correct operating current for the bolometer used.

The instrument is now ready for use.

THEORY OF OPERATION

2.1 RF Attenuation.

The unknown attenuation of a microwave component may be determined very conveniently and accurately by a method of measurement known as the "Ballantine Voltmeter Method." (For a discussion of this method, see pages 806-808 and 838-841 of Volume 11 of the Radiation Lab. Series, "Technique of Microwave Measurements" by C. G. Montgomery.)

2.2 Antenna Patterns.

For detailed information concerning the methods used for antenna measurements, see pages 599 and 602-606 of Volume 12 of Radiation Lab. Series, "Microwave Antenna Theory and Design" by S. Silver.

2.3 Impedance Measurements.

Standing Wave Ratio measurements can also be made by using the BA-5 in conjunction with a slotted line and suitable detector. If a crystal is used, the dc bias must be switched off. The theory is covered in many standard text books, such as, e.g., pages 122, 145 and 153 of "Electronic Measurements" by Terman and Pettit (2nd Edition, 1952, McGraw-Hill).

2.4 Equipment Required.

The method used and equipment required are described in the BA-1 Bolometer Preamplifier Instruction Manual IM#1. However, since the BA-5 Bolometer Amplifier and Attenuation Calibrator combines certain functions of the equipment listed therein, the information is repeated with necessary modifications for the BA-5. A list of suitable equipment is presented in Table 1.

2.5 A Single Channel Insertion Loss Test Set-Up is shown in the block diagram B-114-103. This set-up is capable of measuring R.F. losses to 30 db in one step with an accuracy of .1 db. The maximum peak power required for the barretter is 200 microwatts.

2.6 A Two-Channel Insertion Loss Test Set-Up is shown in the block diagram B-114-121. The range covered is 0-20 db. The 2nd channel is a reference or "bucking" channel which compensates automatically for random amplitude variations of the R.F. source.

For variations of the R.F. source of plus or minus .2 db, insertion loss measurements between 0 and 10 db can readily be repeated to .01 db and measurements between 10 and 20 db are repeated within .01 and .02 db.

Because the insertion loss depends upon load and source impedance, both have to be well defined in order to make precise insertion loss measurements, then accuracies of .02 db/10 db can be realized. Errors due to reflection interactions and crosstalk are discussed on page 824 and page 830 of "Technique of Microwave Measurements" by Montgomery (Volume 11, M.I.T. Radiation Lab. Series, McGraw-Hill).

Additional theoretical considerations on deviation of barretters from square law and influence of noise are contained in our Engineering Notes 1, 2 and 3 which are attached.

2.7 General Notes Concerning Associated Equipment.

2.7.1 RF Source -- The power source should be a modulated r-f generator with at least 20 milliwatts peak output. It must be 100% square wave modulated from an accurate and stable AF source at 1,000 cps. Both the r-f oscillator and the audio oscillator must be stable in frequency and amplitude, and possess stable duty cycles. The Weinschel Modulated R-F Sources MS-1, MS-2 or MS-3 (50-250, 250-920, 900-2000 mc) are suitable. Some means of controlling the output power level are necessary. This control is generally an integral part of the R.F. source. The square wave modulator may or may not be part of the generator, but it is essential for accurate results that the generator be 100% modulated at the audio frequency to which the amplifier is tuned (1000 cps plus or minus $\frac{1}{2}$%).

When making insertion loss measurements, the device under test is inserted usually between a source and load, both of which have the same impedance, matching the characteristic impedance of the line. If the generator's impedance does not match the line impedance, but adequate output is available, the simplest method of providing the proper source impedance is to connect a 20 db pad having a low VSWR in series with the output of the generator. This will "mask" the generator's output impedance and the output terminals of the pad will represent a source impedance which is close to the desired line impedance.

If the generator output level is not sufficient to compensate for the added loss of the 20 db pad, tuning stubs can be used to effect the desired match.

2.7.2 Slotted Line. The slotted line should have a residual VSWR of 1.04 or better, preferably on the order of 1.02. This is necessary in order to adjust the bolometer mount to match the characteristic impedance of the system. It is good practice to use the VSWR indicator which is the companion instrument to the slotted line used. If the indicated VSWR figure is to be correct, care must be exercised to operate the detector of the slotted line in the square law region. Extreme care must be observed in the selection and use of adapters and connectors from

the load side of the slotted line to the unknown component, the tuning stubs and the bolometer mount. The component whose attenuation is to be measured is inserted in the system between the slotted line and the combination of stub tuner and bolometer mount.

2.7.3 Bolometer Mount. Most of the bolometer mounts recommended do not require a stub tuner ahead in the system for a d-c return. However, if lower VSWR readings are desired, it is important to include stub matching ahead of the bolometer mount. This method of measuring attenuation is limited in r-f frequency range only by the bolometer mount. It can be accomplished at any frequency at which it is practical, and feasible, to match the bolometer mount to the transmission system. The practical frequency range can be considered as extending from 20 mc to 40 kmc. Some of the commercially available bolometer mounts are broad band and can be used over frequency ranges of five or ten to one; examples are the PRD 623 (20-1000 mc) and 627A (500-10,000 mc). These mounts normally do not require matching stubs for good results. Other bolometer mounts require the stubs not only for matching, but also for a d-c return; such a mount is, e.g., the PRD 613. Below 20 mc, it is necessary to increase the coupling and bypass capacities of most commercially available bolometer mounts.

2.7.4 Tuning Stubs. At these lower frequencies, tuning stubs become bulky; therefore, lumped tuning or resistive masking with a good 10 db or 20 db pad is more practical. Around 100 mc, a practical double stub tuner can be constructed from General Radio stubs type 874-D50 and type 874-LA Adjustable Line, Tees 874-T and Adapters 874-QNP and 874-QNJ. For convenience of operation, the 2 stubs should be 1/8 wavelength apart and each stub must cover a travel 1/2 wavelength long. All parts should be clamped solid to a support, e.g., a board. In some instances, the length of these stubs must be increased with type 874-L Line Elements.

The cost of bolometer elements is quite high and they are a nuisance to replace; therefore, every effort has been made in the design of the BA-5 to prevent burnout of bolometer elements by DC surge currents. For example, with a BA-5, it is possible to disconnect and reconnect the bolometer without any danger of burning out the element. It is not necessary to shut off the bias current when so doing. The DC meter indicates the bolometer bias current, and permits adjustment to the correct level.

2.7.5 Barretter Element. The barretter in the bolometer mount is biased by a d-c current furnished by the Amplifier. This current is indicated by the meter on the amplifier marked "Bolometer Current." It is controlled by a potentiometer on the chassis behind the meter. The bias current should always be within the

limits specified by the manufacturer. This current is generally 4.5 or 8.75 milliamperes respectively. When biased as described, the barretter will operate in the square law region (for amount of error, see attached Drawing B-114-33) providing the maximum r-f power level applied is also taken into consideration. The maximum r-f power level for square law detection of a PRD 610A barretter is about 200 microwatts peak power. In a true square law detector, the change in resistance is proportional to r-f power; therefore, the audio voltage developed across the barretter is proportional to r-f power. In other words, for any change in r-f power expressed in db, the resulting change of audio voltage expressed in db is twice the r-f change. Theoretical and experimental studies show that under the above conditions the error in attenuation measurements due to deviation from square law does not exceed .02 db R.F. (actually, it is close to .01 db R.F.)

2.8 Bias Current.

The bolometer bias current should always be within the limits specified by the manufacturer. The barretter in the bolometer mount is biased by DC obtained from the B_+ supply via voltage dropping and stabilizing network. The bias current is indicated by the 0-10 ma meter M1. Because bolometer bias adjustments are not constantly being made, and to avoid the risk of involuntary misadjustment of the bias control, the potentiometer P-2 is sub-panel mounted, and adjustment is made by means of a screwdriver through a hole in the panel. The current range of 0-10 ma accommodates most types of bolometers.

Since the bias current is obtained from a 300 v dc source by means of dropping resistors, it is practically a constant current over a wide range of bolometer resistance.

2.9 Gain.

A 30 db r-f power change produces a resulting change in audio voltage of 60 db. Since the audio voltage at 200 microwatts peak r-f power is only about 2+ millivolts RMS of the 1000 cps fundamental across the barretter, an r-f drop in power level of 30 db to .2 microwatts produces only 2 microvolts of audio across the barretter. A 35 db R.F. loss correspondingly drops the A.F. level 70 db. In order to measure R.F. loss to the maximum signal/noise ratio in the range of 0-35 db, it is necessary to have a gain adjustment covering 0-70 db A.F. In the BA-5, the amplifier operates with fixed gain during a measurement. A maximum gain of 50×10^6 is available in the amplifier to provide full output meter deflection with an input as low as 2 microvolts. The output meter is always operated at full scale deflection, hence meter calibration errors do not enter into the results obtained.

2.10 Standard A.F. Attenuator.

An accurate stepattenuator is located between the bolometer input terminal and the input to the bolometer amplifier. Its accuracy at 1000 cps is such that for R.F. measurements to 30 db, it may introduce a maximum error of .01 db (R.F.).

In making a measurement, more attenuation than the expected loss in the attenuator under test is inserted in the reference audio attenuator which immediately f dlows the bolometer output and precedes the amplifier. The amplifier gain is adjusted for full scale meter deflection. When the unknown attenuator is inserted, the loss in the reference audio attenuator is reduced until the meter again reads full scale. The only change made in the settings of the instrument is the adjustment of the reference audio attenuator. Thus it will be seen that the overall accuracy of the BA-5 itself depends entirely on the accuracy of the calibrated attenuator.

2.11 Noise.

The minimum audio voltage when measuring 30 db attenuation under the conditions outlined is 2 microvolts. In order to measure this low voltage level accurately, the noise contributed by the amplifier should not exceed 1%, since the error in the indication of the VTVM due to the presence of noise is of the same order. The noise level of the BA-5 Amplifier is about 25-30 millimicrovolts referred to the input, when operated with a shielded 200 ohm wirewound resistor across the input jack.

2.12 Input Impedance.

In order to utilize most of the audio voltage developed across the barretter, the amplifier should have a high input impedance when compared with the normal barretter resistance of 200 ohms. Changes in the barretter resistance caused by changes in r-f level usually result in a change of the voltage division between the barretter and amplifier input, unless both are purposely mismatched in the right direction. For this reason, the input impedance of the BA-5 Amplifier is designed to be 2000 ohms.

2.13 Audio Frequency.

The frequency used for modulation and detection is determined by two factors. In order to be relatively free from noise, the frequency should be either low or high compared with 60 cps, 60 cps harmonics and microphonics. In order to utilize standard transformer designs, a higher frequency was chosen. However, due to the thermal time constant of the barretter, (for details, see Weinschel Engineering Notes No. 3, April 1955, on "Thermal Time Constant of a Bolometer" which are part of this Manual) its demodulation efficiency and audio output decreases with increasing frequency. The PRD 610A barretter, for example, has a time constant of about 120 microseconds. Its audio output at 400 cps is reduced by 2 db (audio, i.e., 1 db R.F.) and at 1000 cps by 4 db (audio, i.e., 2 db R.F.) when compared with 100 cps. Therefore, as a compromise, we operate at 1000 cps.

2.14 Extending The 30 db Range.

2.14.1 By R.F. Substitution: Power ratio measurements exceeding the normal 30 db range of the equipment described, can be made in 2 steps by calibrating a good, low VSWR 30 db pad. With this pad, together with the requirements of raising the r-f level from the 200 microwatts peak power (square wave modulated) to 200 milliwatts peak power, it becomes practical to measure a 60 db loss. It must be remembered, however, that whatever error exists in the calibration of the 30 db pad must be added to the system error. We can supply a 30 db, 1 watt pad calibrated to an accuracy of .1 db. Using such pads, the total error of a single channel system for a 60 db measurement as described above is in order of .2 db.

The source impedance should be adjusted to represent the proper impedance Zo by stubs.

This can also be accomplished by using the output terminal of a directional coupler of Zo as a source while monitoring the incident power, e.g., by means of a H.P. #415 VSWR Meter. Adjust manually the output of the R.F. source to maintain a constant level of incident power both when the range extending pad as well as when the unknown attenuator is connected to the coupler. This is equivalent to operation from an R.F. source of impedance Zo. In both cases, the VSWR of the range extending pad becomes unimportant.

2.14.2 By Using Barretter in Non-Linear Range: When extending the range of a single step measurement by using more than 200 microwatts peak power, it is generally desirable to take into consideration the error introduced by the barretter's deviation from true square law response. This error may be approximated and corrected for, by the application of corrections based on the equations outlined in the attached drawing B-114-33. A typical computation of the resultant error is shown below for the PRD 610-A:

Exponent of Resistance of Barretter = n = .9
P_1 = D.C. Bias Power = 4 milliwatts (4.5 MA, 200 ohms)
ΔP = Maximum R.F. Power = 200 microwatts peak
$(n-1)\frac{\Delta P}{P_1} = .005$

Consulting the graphs in Drawing B-114-33, we find maximum error .011 db (R.F.)

This error due to the deviation of the barretter from square law results in a lower measured attenuation value. For correction, it ought to be added to the measured value.

Subject to the limitations due to burnout of the barretter (e.g., 8 milliwatts for the PRD 610-A), one can operate at a higher R.F. power level and make a correction for the deviation. A level

of 2 milliwatts peak corresponds for a PRD 610-A to about 20 millivolts, 1,000 cps at the input of the BA-5. According to the above mentioned graph, a measurement of 10 db R.F. loss requires a plus .095 db correction. A measurement of an R.F. loss exceeding 20 db requires a correction of plus .11 db. Using this correction, it is simple to measure a 40 db loss, in one step, by adding .11 db to the measured attenuation.

By measuring, e.g., the same 20 db pad at a maximum R.F. level of 200 microwatts and 2 milliwatts, one may determine experimentally the error introduced by the barretter.

2.15 Attenuation Calibrations When Making A.F. Measurements.

The attenuator is marked in db (R.F.) changes produced in the output of a square law detector whose voltage output is proportional to the power input. For this reason, the 1000 cps attenuation is actually twice the (R.F.) attenuation marked on the panel. If the BA-5 is used for 1000 cps measurement or in combination with linear detectors, multiply all attenuator settings and meter indications in decibels by a factor of two.

3.1 To Measure Up To 30 db Attenuation (Single Channel).

To make measurements up to a 30 db power ratio, recommended equipment (see Table 1) should be set up as shown in Drawing B-114-103.

It is important that the equipment selected by appropriate for the frequency range desired. The procedure to follow in making measurements is outlined below:

3.1.1 The r-f generator is tuned to the desired frequency and 100% modulated with a 1000 cps (plus or minus ½%) square wave. The pad or attenuator to be measured should not be in the circuit at this stage of measurement. Adjust the generator to a low output level to avoid the risk of burning out the bolometer element. For good accuracy, do not exceed 200 microwatts peak power (100 microwatt average power) for the PRD 610-A barretter (See Section 2.14.2).

Before switching on the BA-5, set both gain controls to their maximum counterclockwise stop for minimum gain. Rotate the "bolometer current adjustment" to its maximum counterclockwise stop. (This is not necessary if the adjustment has been set previously for the correct bias current). Set all attenuators to zero. Operate the power switch to "on." The red pilot light should light up at once. The output meter should stay at zero. Then adjust the bolometer current to the value recommended by the manufacturer of the barretter; this is, e.g., 4.5 MA for a PRD 610-A. Match roughly the bolometer mount to the transmission line by adjusting the double stub tuner ahead of the bolometer mount, which results in a maximum reading of the BA-5 meter. However, with the BA-5 gain at minimum and all BA-5 attenuators set to zero, keep the R.F. level low enough so that the meter does not exceed a full scale indication. This corresponds to a 1000 cps input signal of about 2.5 millivolts into the BA-5. This is a quick check to ascertain operation in the square law region for all R.F. barretters having time constants below 500 microseconds and bias currents below 10 MADC and a DC resistance between 50 and 200 ohms. Most bolometers presently supplied by F. & R. Machine Works, Narda, P.R.D. and Sperry fall into this category.

For consideration of thermal time constants and frequency response, consult Engineering Note No. 3. For the initial measurements, check the voltage across the bolometer terminal. It has been found most helpful in actual operation to leave a BNC-T connector (e.g., UG-274/U) in the bolometer input

terminal of the BA-5. This permits monitoring of the A.C. voltage at the input. Set it to 2 millivolts at maximum R.F. level before insertion of the unknown attenuator for best results. Next, tune the slotted line probe to the R.F. frequency for maximum indication on the VSWR meter. Adjust the double stub tuner bolometer mount combination for minimum VSWR. This should be close to the residual VSWR of the slotted line. The system is now matched to Zo. It is well to recheck the VSWR again at this power level. It should be the same as before. Set the BA-5 reference attenuator to a value which exceeds the attenuation of the R.F. loss to be measured by at least 1 db. Increase the gain of the amplifier until the output meter reads full scale. Insert the attenuator or pad to be measured between the slotted line and the tuning stubs. Do not change any other adjustments. Decrease the setting of the Audio Stepattenuator until the BA-5 indicator returns to the same reading as before. The difference between the two db settings of the BA-5 Audio attenuator is the unknown insertion loss of the R.F. reference attenuator or pad.

3.2 To Measure Attenuation Greater Than 30 db.

It is practical to measure attenuation in excess of 30 db in the following manner. First, calibrate a 10, 20 or 30 db attenuator using the procedure outlined in Section 3.1. This attenuation, e.g., a 20 db pad, is then left in the r-f system. The BA-5 reference attenuator should be set to about 30 db with the gain adjusted to produce a full scale reading for 200 microwatts R.F. peak power into the bolometer mount. The power level of the r-f generator is increased to produce again a full scale reading. This increases the signal generator level by the amount of the "range extending" attenuator. Without touching any controls, insert the unknown attenuator, and remove the calibrated 20 db pad.

Now remove attenuation on the BA-5 reference attenuator until the BA-5 output meter again reads full scale. The sum of the attenuation removed plus the calibrated loss of the nominal 20 db pad equals the insertion loss of the unknown pad. Using this system, it is possible to measure up to 60 db of attenuation, providing the R.F. source is capable of supplying 200 milliwatts peak power from source impedance Zo.

3.3 Noise.

The output meter can be switched to a long or a short time constant to facilitate readings.

3.4 Measurement of VSWR.

The BA-5 is also designed for use in making VSWR measurements and reading VSWR values directly on the BA-5 output meter, which is calibrated accordingly. The BA-5 should be used with the appropriate crystal or bolometer probe. If a crystal probe is used, it is imperative to remember to switch off the bias voltage, otherwise, the crystal might be damaged. Also be sure that the probe coupling is adjusted for operation in the square law region.

3.4 Direct Measurement of R.F. Losses to 1 db With Accuracy Of .01 db.

Set up equipment to comply with blockdiagram on Dwg. B-114-121. For operation refer to Instruction Manual IM#3 on ND-1.

3.5 Measurement of A.F. Losses 0-20 db With Accuracy of .02 db/10 db.

Set up equipment as in the preceding paragraph and use Instruction Manual IM#3 on ND-1.

4.1 Introduction.

The Model BA-5 Amplifier and Attenuation Calibrator has been designed to require a minimum of maintenance. However, should trouble develop and service become necessary, the usual procedure for maintenance of vacuum tube amplifiers applies.

4.2 General Maintenance.

4.2.1 Tubes: The checks usually made to determine the condition of vacuum tubes will normally suffice. However, special selection of the tube V7(5814) may be required if the noise level is high or if the amplifier has a tendency to be microphonic. Generally, three out of five new tubes tested will be good. The test described in 4.3.1 below should be made if tube trouble of this nature is experienced.

4.2.2 Components: Resistance, capacity and voltage values are shown in the attached drawing D-114-122. Voltage measurements were made using a 20,000 ohms per volt meter. The "Twin-Tee" plug-in units are held to close tolerances and adjusted during manufacture. It is not advisable to attempt any repair of this component. Should the "Twin-Tee" units become defective, they should be returned to us for repair and adjustment. Never inter-change the two twin-tee plug-in networks.

4.2.3 Oscillation: With extremely high gain amplifiers, conditions of oscillation are occasionally encountered. In the BA-5, oscillation is indicated by a steady reading on the output meter, with or without a connection to the input jack. If the amplifier should go into an unstable condition, it is probably caused by a high resistance ground, or a broken ground connection.

4.2.4 B_+ Supply: The B_+ voltage should be approximately 330 volts to the amplifier, and 300 volts to the special constant current bolometer bias supply network. The power supply circuit employs both current and voltage regulation.

4.2.5 Input Termination: UNDER NO CIRCUMSTANCES MUST PRE-SET POTENTIOMETER P-5 BE ADJUSTED. This is adjusted and locked at the factory during final testing to provide a correct 2000 ohm termination for the precision input attenuator. Any change of its setting will cause errors in the calibration of the attenuator and necessitate its being returned to the factory for re-setting.

4.3 Performance Tests.

If the BA-5 Amplifier is suspected of faulty performance, several simple tests may be made which will establish its agreement with the major original specifications. These tests are shown below.

4.3.1 Noise: In use, the BA-5 noise levels should be always the same during a measurement. However, if erratic results are being

obtained, and noise is suspected, the following tests can be made. A shielded 200 ohm resistor is plugged into the input, and the bolometer bias control adjusted to read 9 ma on the meter.The output meter time constant switch should be in the "short" position. The BA-5 should not have more than 30 millimicrovolts of equivalent noise at the input. This is indicated by a reading of between 3 and 4 on the VSWR scale with all attenuation "out", and gain controls set to maximum.

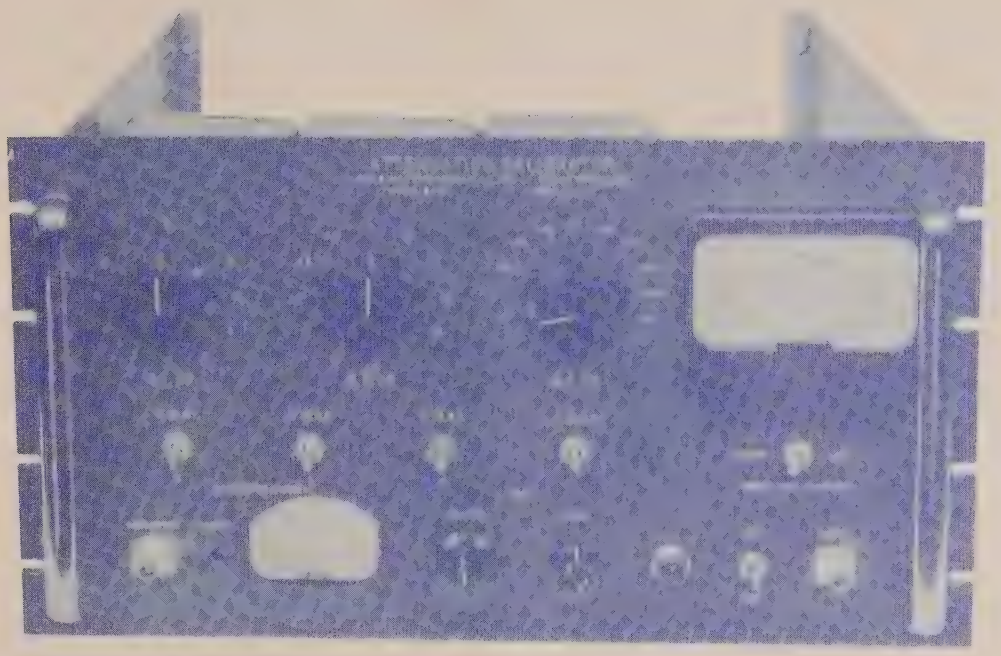
If a higher value of noise is indicated, set the coarse gain control to minimum and observe the output meter. If the reading is substantially unchanged, it indicates that the noise is occurring in the stages after the coarse gain control.

If reducing coarse gain decreases the noise reading, the noise may be generated in the bolometer circuit.

Tubes should be checked for noise and microphonics, and in particular the type 5814 tubes should be carefully selected for low noise by trying a number of tubes in the amplifier and measuring the noise level.

4.3.2 Audio Filter: The 1 KC filter is a permanent part of the BA-5 circuit and is not adjustable. Its bandwidth is designed to be 40 cps or less. To check this, connect a tunable 1 KC oscillator to the input, and adjust the gain controls and attenuator, to give approximately a half scale deflection with an input of 1 KC. Vary the frequency of the oscillator and note the meter reading. It should be constant through the bandpass of the amplifier. Note the highest and lowest frequency at which the output commences to fall off. These points must not fall more than 5 cps on each side of 1 KC.

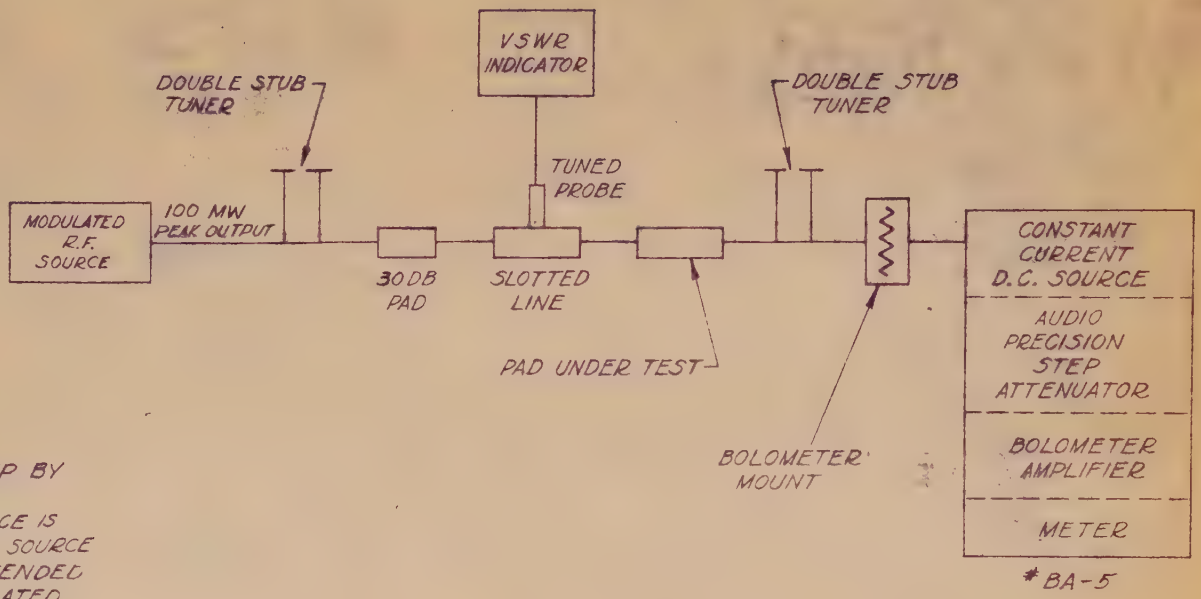
If the audio filter selectivity appears to be incorrect, it is best to ship the instrument back to the factory for repair, rather than attempt to substitute for the matched components in the BA-5 filters.



P - 46
Attenuation Calibrator
Model BA-5

TOLERANCES
ALL 3 PLACE DECIMAL DIMS. ± .003
ALL 2 PLACE DECIMAL DIMS. ± .01
ALL FRACTIONAL DIMS. ± 1/64
ALL ANGLES ± ½°
UNLESS OTHERWISE SPECIFIED

LIST OF MODULATED R.F. SOURCES	
MODEL NO.	FREQUENCY
MS-1	50-250 MC
MS-2	250-920 MC
MS-3	900-2000 MC
MS-4	2000-4000 MC



NOTES:
1- 30 DB CAN BE MEASURED IN ONE STEP BY USING THE BA-5.
2- IF THE GENERATOR OUTPUT IMPEDANCE IS ADJUSTED TO PRESENT THE PROPER SOURCE IMPEDANCE, THIS RANGE MAY BE EXTENDED TO 60 DB BY THE USE OF A CALIBRATED 30 DB PAD APPLYING THE R.F. SUBSTITUTION METHOD.

INSERTION LOSS TEST SET NO. 1, GROUP 2
0-30 DB (0-60 DB)

REVISIONS
REV. LTR. | DESCRIPTION | DATE | APPVD
A | REVISED | 5-1-64 | [illegible]

USED ON	ASSY. DRWG.

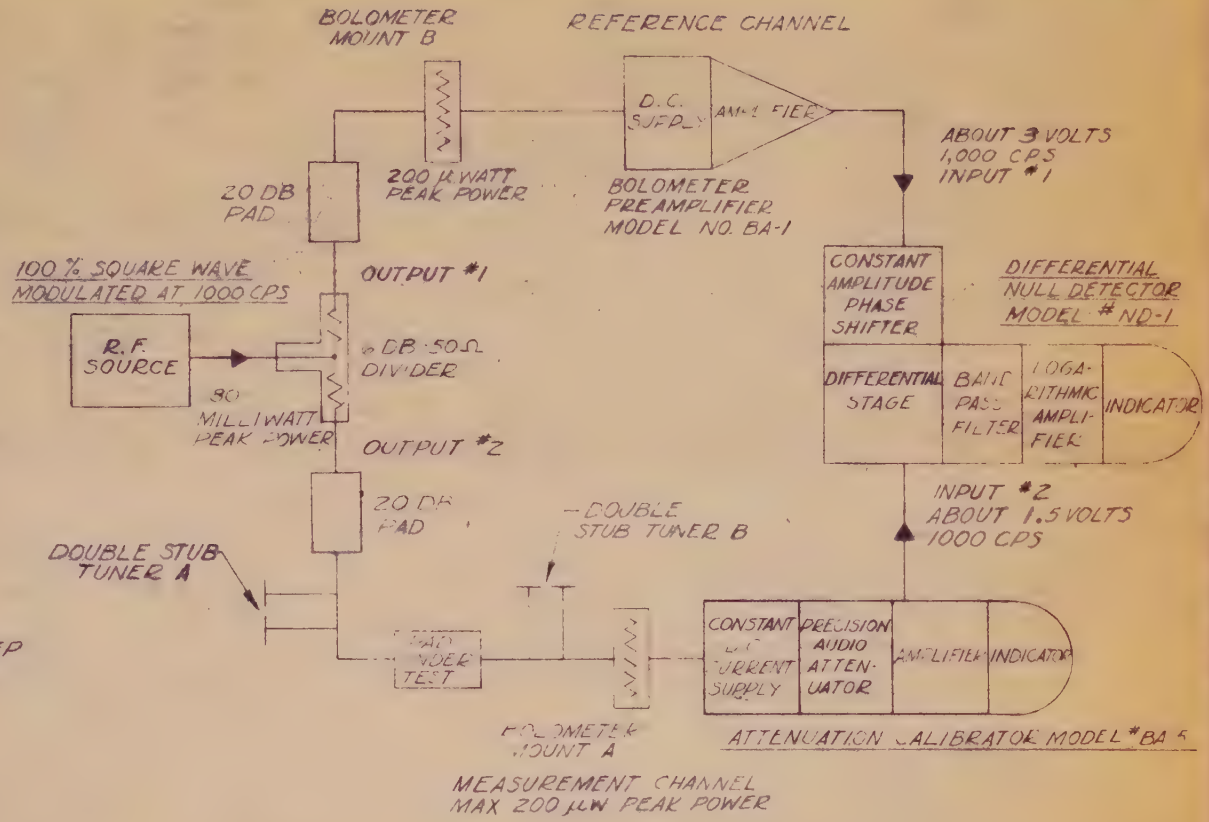
Weinschel Engineering Co.
SILVER SPRING, MD.

INSERTION LOSS TEST SET NO. 1, GROUP 2 (0-30DB) (0-60DB)

MATERIAL
FINISH
D 7/29/54 D.A.C. | C | T | 22
SCALE
B 114-103 | A

TOLERANCES
ALL 3 PLACE DECIMAL DIMS. ± .003
ALL 2 PLACE DECIMAL DIMS. ± .01
ALL FRACTIONAL DIMS. ± 1/64
ALL ANGLES ± ½°
UNLESS OTHERWISE SPECIFIED

LIST OF MODULATED R.F. SOURCES	
MODEL NO.	FREQUENCY
MS-1	30-250 MC
MS-2	250-920 MC
MS-3	900-2000 MC
MS-4	2000-4000 MC



NOTE:
1- 0-20 DB CAN BE MEASURED IN ONE STEP BY USING BA-1, BA-5, AND ND-1.

REVISIONS		
A	REVISED	7-11-55

USED ON	ASSY DRWG.

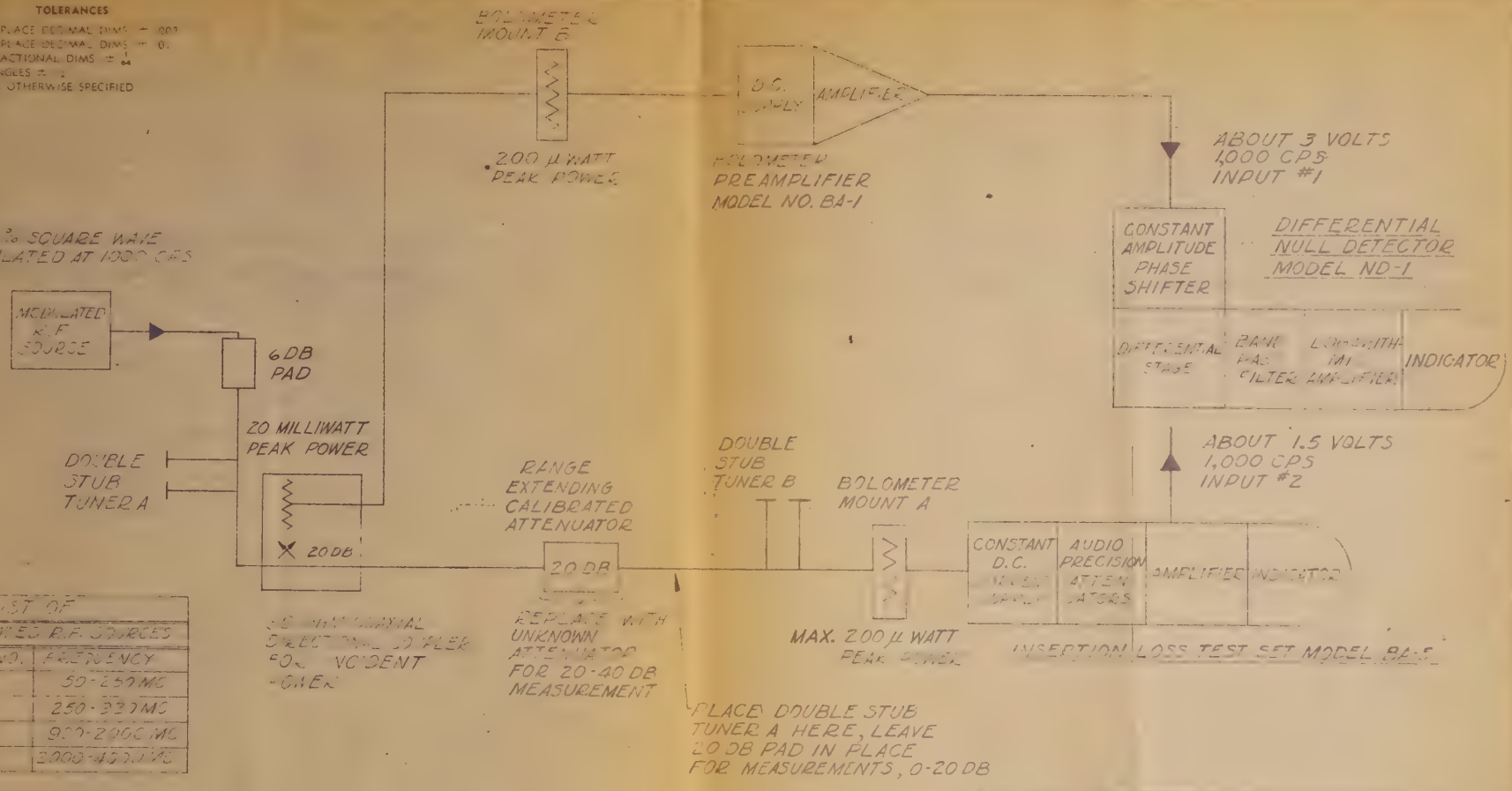
Weinschel Engineering Co.
SILVER SPRING, MD.

INSERTION LOSS TEST SET NO.3 0-20DB GROUP I

MATERIAL
FINISH
D C SCALE
T A B 11-1-121 A

8114-121 A

TOLERANCES
ALL 3 PLACE DECIMAL DIMS ± .002
ALL 2 PLACE DECIMAL DIMS ± .01
ALL FRACTIONAL DIMS ± 1/64
ALL ANGLES ± 1°
UNLESS OTHERWISE SPECIFIED


BOLOMETER MOUNT B
200 μ WATT PEAK POWER
D.C. SUPPLY
AMPLIFIER
BOLOMETER PREAMPLIFIER MODEL NO. BA-1
ABOUT 3 VOLTS 1,000 CPS INPUT #1
CONSTANT AMPLITUDE PHASE SHIFTER
DIFFERENTIAL NULL DETECTOR MODEL ND-1
DIFFERENTIAL STAGE
BAND PASS FILTER
LOCK-SMITH METER AMPLIFIER
INDICATOR
100 % SQUARE WAVE MODULATED AT 1000 CPS
MODULATED R.F. SOURCE
6DB PAD
20 MILLIWATT PEAK POWER
DOUBLE STUB TUNER A
X 20DB
RANGE EXTENDING CALIBRATED ATTENUATOR
20 DB
DOUBLE STUB TUNER B
BOLOMETER MOUNT A
MAX. 200 μ WATT PEAK POWER
ABOUT 1.5 VOLTS 1,000 CPS INPUT #2
CONSTANT D.C. POWER SUPPLY
AUDIO PRECISION ATTENUATORS
AMPLIFIER
INDICATOR
INSERTION LOSS TEST SET MODEL BA-1
REPLACE WITH UNKNOWN ATTENUATOR FOR 20-40 DB MEASUREMENT
PLACE DOUBLE STUB TUNER A HERE, LEAVE 20 DB PAD IN PLACE FOR MEASUREMENTS, 0-20 DB

LIST OF MODULATED R.F. SOURCES	
MODEL NO.	FREQUENCY
MS-1	50-250 MC
MS-2	250-930 MC
MS-3	900-2000 MC
MS-4	2000-4200 MC

NOTE:
1- 0-20 DB CAN BE MEASURED IN ONE STEP
2- 0-40 DB CAN BE MEASURED BY THE USE OF A CALIBRATED 20 DB PAD APPLYING THE R.F. SUBSTITUTION METHOD.

REVISIONS

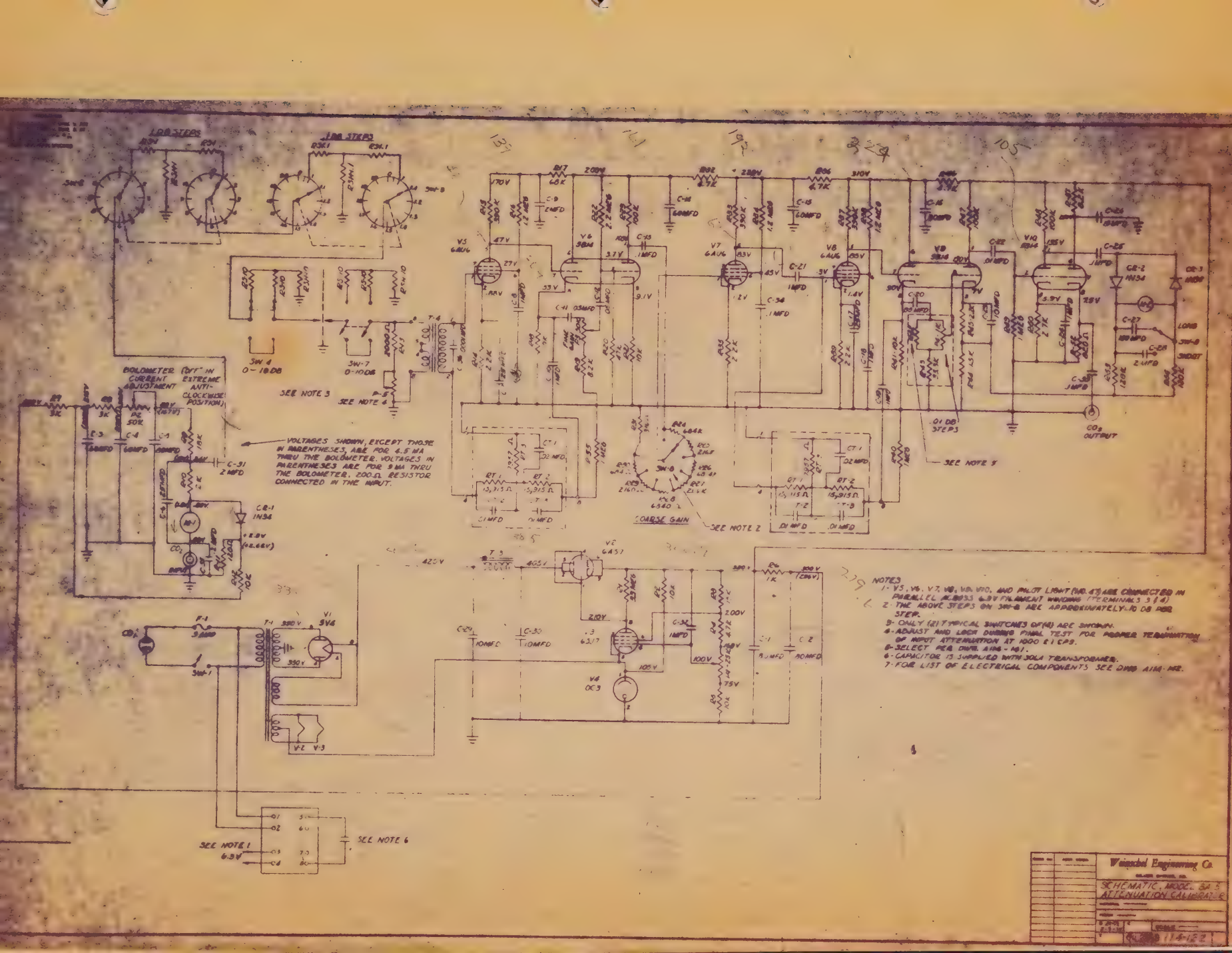
A	REVISED 1/10/56	D.A.C.

USED ON	ASSY DRWG

Weinschel Engineering Co.
SILVER SPRING, MD.

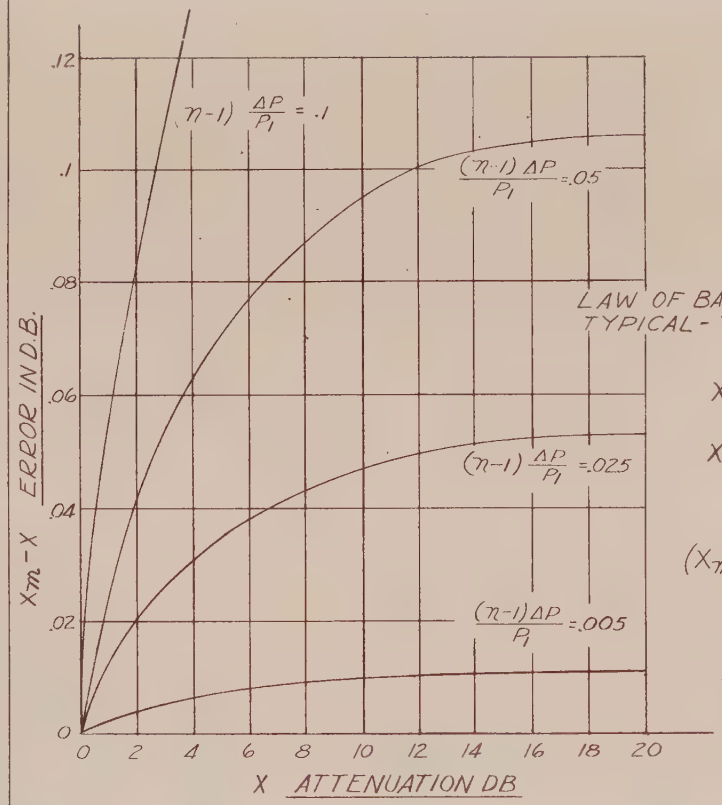
INSERTION LOSS TEST SET
MODEL BA-1 (0-40 DB) SEC. NO. 3

MATERIAL
FINISH
SCALE
B 114-169 A


VOLTAGES SHOWN, EXCEPT THOSE IN PARENTHESES, ARE FOR 4.5 MA THRU THE BOLOMETER. VOLTAGES IN PARENTHESES ARE FOR 9 MA THRU THE BOLOMETER. 200 Ω RESISTOR CONNECTED IN THE INPUT.
BOLOMETER CURRENT ADJUSTMENT
SEE NOTE 1
SEE NOTE 2
SEE NOTE 3
SEE NOTE 4
SEE NOTE 5
SEE NOTE 6
COARSE GAIN
01 DB STEPS
CO2 OUTPUT
NOTES
1- V5, V6, V7, V8, V9, V10, AND PILOT LIGHT (NO. 47) ARE CONNECTED IN PARALLEL ACROSS 6.3V FILAMENT WINDING (TERMINALS 3 & 4).
2- THE ABOVE STEPS ON SW-8 ARE APPROXIMATELY .10 DB PER STEP.
3- ONLY (2) TYPICAL SWITCHES OF (4) ARE SHOWN.
4- ADJUST AND LOCK DURING FINAL TEST FOR PROPER TERMINATION OF INPUT ATTENUATION AT 1000 & 1 CPS.
5- SELECT PER DWG. A114-161.
6- CAPACITOR IS SUPPLIED WITH SOLA TRANSFORMER.
7- FOR LIST OF ELECTRICAL COMPONENTS SEE DWG A114-162.
Weinschel Engineering Co.
SCHEMATIC, MODEL 8A-3 ATTENUATION CALIBRATOR
114-152

TOLERANCES

ALL 3 PLACE DECIMAL DIMS. ± .003
ALL 2 PLACE DECIMAL DIMS. ± .01
ALL FRACTIONAL DIMS. ± 1/64
ALL ANGLES ± ½°
UNLESS OTHERWISE SPECIFIED



X = ATTENUATION IN DB
ΔP = MAX. R.F. POWER (PEAK) AT 0 DB
P_1 = D.C. BIAS POWER
$\Delta P'$ = R.F. POWER (PEAK) WITH ATTENUATOR INSERTED
X_m = MEASURED ATTENUATION IN DB BY ASSUMING PERFECT SQUARE LAW ($n = 1.0$)

LAW OF BARRETER - $R = R_0 + JP^n$
TYPICAL - $n = .9$ FOR SPERRY #821

$$X = 10 \log_{10} \frac{\Delta P}{\Delta P'}$$

$$X_m = 10 \log_{10} \left[\frac{\Delta P}{\Delta P'} + \frac{(n-1)\Delta P(\Delta P - \Delta P')}{2 P_1 \Delta P'} \right]$$

$$(X_m - X) = 10 \log_{10} \left[1 + \frac{(n-1)\Delta P}{2 P_1} \left(1 - 10^{-X/10}\right) \right]$$

ERROR IN ATTENUATION MEASUREMENT DUE TO DEPARTURE OF BARRETTER FROM SQUARE LAW

SLIDE NO. 6

REVISIONS		
A	REVISED - 1-13-55	DAS

USED ON	ASSY. DRWG.

Weinschel Engineering Co.
SILVER SPRING, MD.

ERROR IN ATTENUATION MEASUREMENT DUE TO DEPARTURE OF BARRETTER FROM SQUARE LAW

MATERIAL
FINISH

D 6-15-54 WKS | C | T | A | SCALE | B114-33 | A

ENGINEERING NOTES NO. 1
February 1955

PRECISE INSERTION LOSS MEASUREMENTS USING IMPERFECT SQUARE LAW DETECTORS AND ACCURACY LIMITATIONS DUE TO NOISE

by
Gunther U. Sorger
and
Bruno O. Weinschel
Weinschel Engineering Co. Inc.
Kensington, Md.

Summary

Inaccuracies of insertion loss measurements caused by deviation of a square law detector from square law sets a practical upper limit on the r.f. power which can be applied to a bolometer.

The detection of the audio modulation after extraction from the carrier by a bolometer is limited by the behaviour of the second detector in the presence of noise and sets a lower limit.

Since square law and synchronous detectors are superior to linear detectors at a low "signal to noise ratio", their behaviour is studied in more detail. It is shown that the bandwidth of the amplifier ahead of the second detector is not the controlling factor for the accuracy of the measurement. The concept of "ratio of amplifier bandwidth to indicator bandwidth" is introduced. Graphs are given for the indicator fluctuations as a function of this concept and of the "signal to noise ratio" at the amplifier output for both synchronous and square law detectors.

A graph is furnished showing the range of insertion loss measurement as a function of the accuracy desired based upon the accuracy limitations at the high and low power end.

Introduction

A very convenient method of measuring an R.F. insertion loss is the use of a square wave modulated R.F. carrier, a square law detector as, e.g., a barretter and means to measure the resulting audio voltage.[1,2,3]

This method has been used between 20 Mc and 40,000 Mc where suitable bolometer mounts and barretters are available for proper termination of coaxial and waveguide transmission lines.

Insertion Loss Test Set-Ups

A typical test set-up is shown in Figure 1.

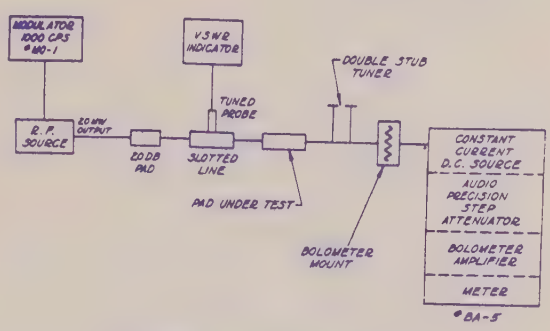
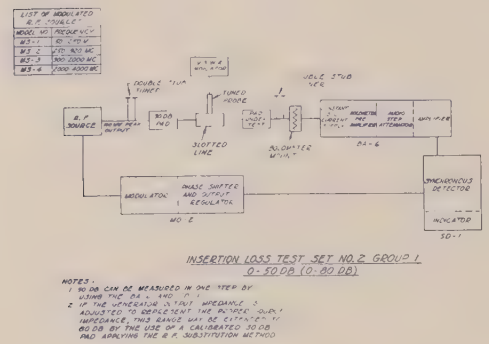


Figure 1

The R.F. source is modulated by a 1000 CPS square wave. The modulated R.F. signal is applied to the bolometer mount which is tuned to properly terminate the transmission line.

The barretter converts the audio modulation of the R.F. carrier to an A.F. voltage. This voltage is amplified and produces by means of a linear detector a meter deflection. The attenuator pad under test, e.g., a 10 db pad is now inserted ahead of the bolometer mount. Because the A.F. voltage developed across the barretter is proportional to the R.F. power, a 10 db reduction in the R.F. level causes a 20 db reduction in the A.F. level. In order to become independent of the amplifier non-linearity and meter calibration, an audio precision attenuator is inserted between the barretter and the amplifier. For a decrease of, e.g., 10 db in the R.F. level, 20 db of audio attenuation is removed to maintain a constant A.F. level at the amplifier input.

The lower power limit of these measurements is determined by noise. A set-up which permits an extension of the lower limit by operating down to a 1:1 signal to noise ratio by means of a synchronous detector is shown in Figure 2.



INSERTION LOSS TEST SET NO.2 GROUP I
0-50 DB (0-80 DB)

NOTES:
1. 90 DB CAN BE MEASURED IN ONE STEP BY USING THE BA-1 AND R.F. ATTENUATOR.
2. IF THE GENERATOR'S OUTPUT IMPEDANCE IS ADJUSTED TO REPRESENT THE PROPER QUARTZ IMPEDANCE, THIS RANGE MAY BE EXTENDED TO 60 DB BY THE USE OF A CALIBRATED 30 DB PAD APPLYING THE R.F. SUBSTITUTION METHOD.

Fig. 2

For measuring the insertion loss of low attenuation devices, high accuracies are usually required. Therefore a dual channel method was developed permitting insertion loss measurement up to 20 db with an accuracy of .02 db/10 db. An important feature of this set-up shown below is that RF level changes do not enter into the measurement.

Deviation From Square Law

Barretters deviate from square law because some of their heat losses are caused by convection and radiation in contrast to conduction. This deviation becomes more important as the operating temperature of the barretter increases which of course is caused by an increase of R.F. power dissipated in the element.

If for practical reasons we restrict our considerations to audio frequencies[4] which are sufficiently low to permit the barretter to follow the modulation envelopes, considering its thermal time constant, the peak temperature is a function of the peak power, not the average power. The resulting deviation from square law is then a function of the peak power.

Depending upon the accuracy required, a limitation is placed upon the peak R.F. power which may be dissipated in the barretter. Typical graphs demonstrating the relation between maximum R.F. peak power, bias power and the static characteristic are given in Figure 3. The derivation of these relations are given in the paper "Insertion Loss Test Sets Using Square Law Detectors."[5]

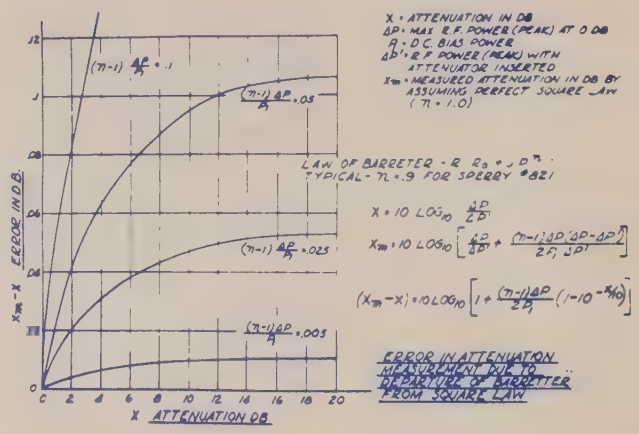

X = ATTENUATION IN DB
ΔP = MAX R.F. POWER (PEAK) AT 0 DB
P = D.C. BIAS POWER
ΔP' = R.F. POWER (PEAK) WITH ATTENUATOR INSERTED
Xm = MEASURED ATTENUATION IN DB BY ASSUMING PERFECT SQUARE LAW (n = 1.0)
LAW OF BARRETTER - R = Ro + I^n
TYPICAL - n = .9 FOR SPERRY #821
X = 10 LOG10 ΔP/ΔP'
(n-1) ΔP/P = .1
(n-1) ΔP/P = .05
(n-1) ΔP/P = .025
(n-1) ΔP/P = .005
ERROR IN DB
X ATTENUATION DB
ERROR IN ATTENUATION MEASUREMENT DUE TO DEVIATION OF BARRETTER FROM SQUARE LAW

Figure 3

For a typical barretter, an error of -.1 db is obtained by using, e.g., a peak power of about 1.5 milliwatts with a 4 milliwatt D.C. bias, and an error of only -.01 db is obtained for a peak power of 150 microwatts. The corresponding 1000 cycle voltages are 25 and 2.5 millivolts R.M.S.

Satisfactory experimental agreement with this theory was obtained in a test set-up shown in Figure 4; both the block diagram as well as the theoretical and experimental deviations are given.

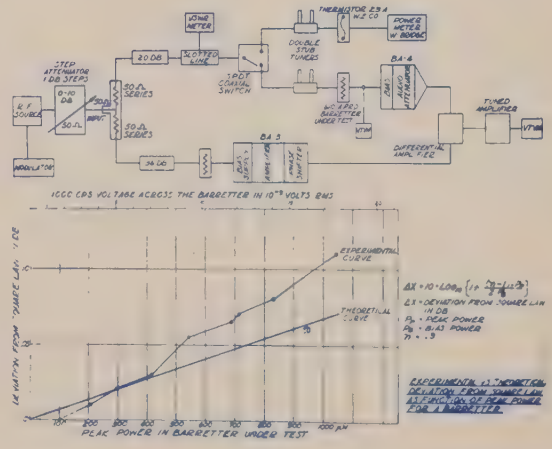


Figure 4

Detector Behaviour in Presence Of Noise

A. Linear Detector

In order to determine the range of insertion loss measurement, we shall now consider the lower limit as determined by noise.

In a linear envelope detector, we experience a phenomenon known as "Signal Suppression by Noise." This becomes very apparent when the signal to noise ratio approaches or is smaller than one.

An envelope detector will suppress a signal equal to noise by 8 db and a signal which is 1/10 noise by 27 db.[6]

For signal to noise ratios below one, the average output of our detector for a constant value of noise will increase by an amount proportional to the square of the signal. The output is also inversely proportional to the first power of the noise, i.e., the larger the noise, the smaller will be the change in output of the detector when the signal is applied. This is the phenomenon which constitutes signal suppression by noise. It is similar to a quadrature combination of the signal and noise; i.e., if we consider for large signal to noise ratios the square root of the sum of the squares of the signal and noise amplitudes, we will find that this square root is relatively independent of the noise and depends directly upon the signal amplitude in a linear way; whereas, for large noise the converse is true, the square root will depend principally upon the magnitude of the noise in a linear way and only very slightly upon the magnitude of the signal amplitude. (See Appendix 1, B for derivation.)

B. Square Law Detector

In a true square law detector the noise energy and signal energy add independently. The contribution of the signal to the average output is independent of the level of noise.

This permits accurate work well below a 1:1 signal to noise ratio, provided that the noise is sufficiently stable. (For derivation, see Appendix 1, C.)

C. Synchronous Detector

The synchronous detector like the square law detector exhibits no signal suppression by noise. Further the average contribution of the noise to the detector output is zero reducing the need to back off the noise voltage.

Fluctuation of Detector Output Due to Noise

Square law and synchronous detectors do not suppress the signal in the presence of noise; however, there are limitations based upon the statistical nature of noise which govern the extension of the range of measurement over the range possible with a linear envelope detector.

Assuming that the noise is stable, its long-time average is constant. This in turn would produce a stable reading in the square law detector which can be backed off. In the synchronous detector, it would result in a stable zero reading in the absence of a synchronous signal and a stable indication in the presence of a synchronous signal.

Usually the noise at the input of the amplifier is white noise. After going through the band pass of the amplifier, it is detected by the synchronous or square law detector.

Let us examine the indicator fluctuations caused by noise either in case of the deflection of a square law detector or the zero reading of a synchronous detector.

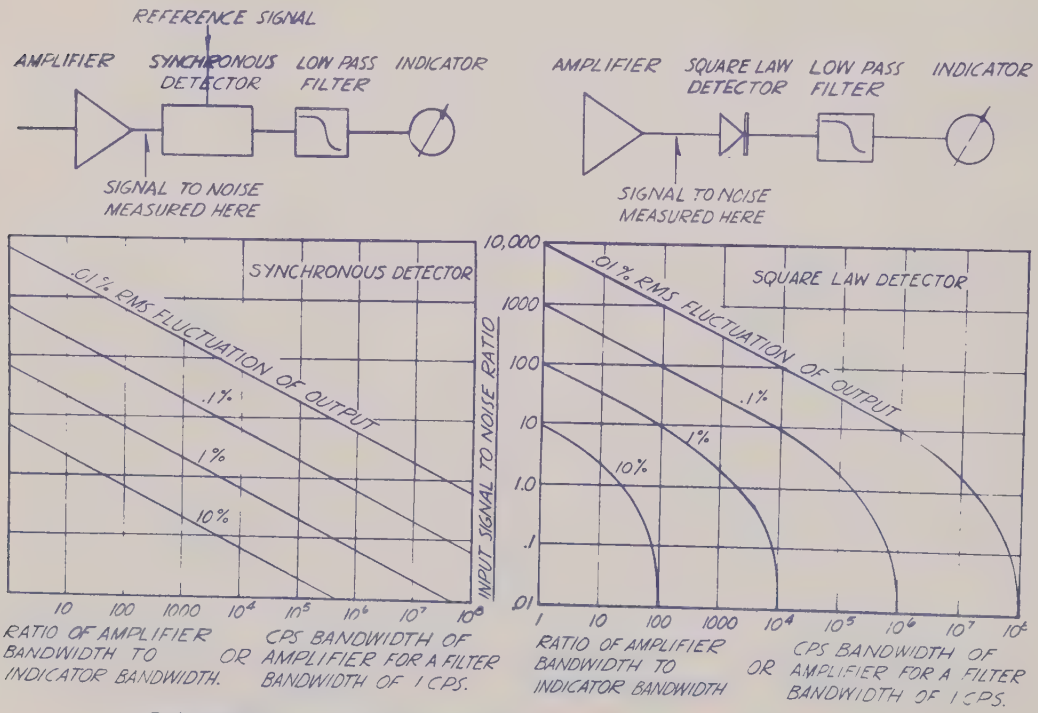
Keeping the input impedance and noise figure of the amplifier fixed, the total noise output power depends upon the amplifier bandwidth. If one reduces, e.g., the bandwidth by a factor of 4 and increases then the gain by 2 to produce at the input of the 2nd detector the same noise output power, the fluctuations of the indicator reading has doubled. (For derivation, see Appendix 2, A.)

The smaller the bandwidth of the amplifier, the greater are the fluctuations of the noise. This can be demonstrated in the following experiment: An electrostatic voltmeter is a square law device. Use a 150 volt F.S. model which is critically damped, having a time constant of $\frac{1}{2}$ second. Connect it to a bolometer amplifier having a bandwidth of 20 CPS. Amplify the noise to produce a 100 volt reading. This reading will then fluctuate about plus or minus 20%. The only way to stabilize this reading is an increase of the mechanical time constant of the indica-

tor.

In case of a second detector, this is equivalent to a reduction of the bandwidth of a low-pass filter between the detector output and the indicator. It becomes apparent that the indicator fluctuations are controlled not by the bandwidth of the amplifier but by the ratio of amplifier bandwidth to indicator bandwidth (See Appendix 2 A,B.)

Figure 5 shows the R.M.S. fluctuation of the indicator for different signal/noise ratios and bandwidth ratios for both synchronous and square law detector.



R.M.S. OUTPUT FLUCTUATIONS AS FUNCTION OF SIGNAL TO NOISE RATIO AND RATIO OF AMPLIFIER BANDWIDTH TO INDICATOR BANDWIDTH

Figure 5

The data of Figure 5 can be used to obtain the lower limit of insertion loss measurements as controlled by noise.

From Appendix 2, we obtain the following relation:

Fluctuation of Output

$$= \frac{\sqrt{2KTRB_F}}{S} \qquad (1)$$

where K Boltzman constant
T Temperature in ^{o}K
R Equivalent noise resistance of amplifier input
B_F Bandwidth of low pass filter
S Signal voltage

and $B_F = \frac{1}{2\tau}$ (2)

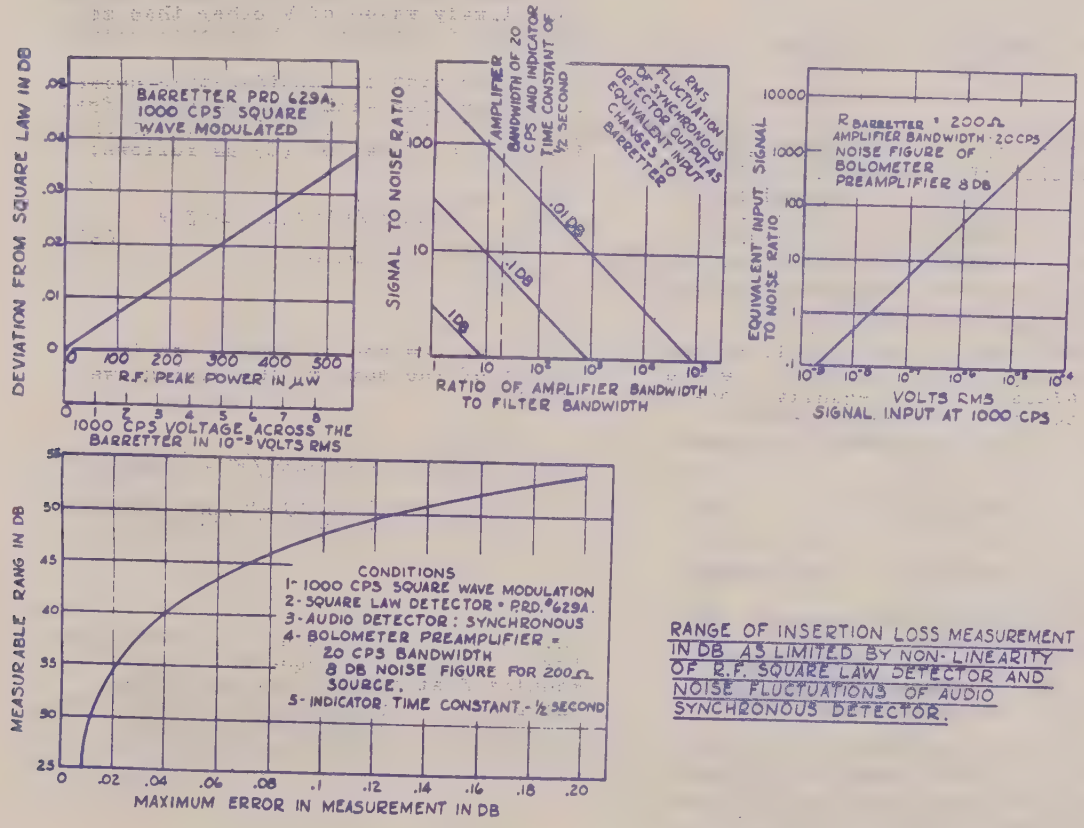
where τ = Time constant of critically damped indicator

It is interesting to note in Equation (1) that the inaccuracy is independent of the bandwidth of the amplifier.

However, it is still advisable to restrict the amplifier bandwidth so that in the presence of noise, the signal is amplified linearly.

These relations are directly applicable to considerations of the bandwidth of I.F. amplifiers and associated second detectors.

Figure 6 shows that a range of about 30 db can be measured with an accuracy of .01 db if half the error is caused by the noise at the low end and half the error is caused by the deviation of the barretter from square law at the high end. Similarly, a range of almost 50 db can be measured with a total error of .1 db.


DEVIATION FROM SQUARE LAW IN DB
BARRETTER PRD 629A, 1000 CPS SQUARE WAVE MODULATED
R.F. PEAK POWER IN μW
1000 CPS VOLTAGE ACROSS THE BARRETTER IN 10⁻⁵ VOLTS RMS
SIGNAL TO NOISE RATIO
AMPLIFIER BANDWIDTH OF 20 CPS AND INDICATOR TIME CONSTANT OF ½ SECOND
RMS FLUCTUATION OF SYNCHRONOUS DETECTOR OUTPUT AS EQUIVALENT INPUT CHANGES TO BARRETTER
.01 DB
.1 DB
1 DB
RATIO OF AMPLIFIER BANDWIDTH TO FILTER BANDWIDTH
EQUIVALENT INPUT SIGNAL TO NOISE RATIO
R BARRETTER = 200Ω
AMPLIFIER BANDWIDTH 20 CPS
NOISE FIGURE OF BOLOMETER PREAMPLIFIER 8 DB
VOLTS RMS SIGNAL INPUT AT 1000 CPS
MEASURABLE RANG IN DB
CONDITIONS
1- 1000 CPS SQUARE WAVE MODULATION
2- SQUARE LAW DETECTOR - PRD. #629A.
3- AUDIO DETECTOR: SYNCHRONOUS
4- BOLOMETER PREAMPLIFIER - 20 CPS BANDWIDTH 8 DB NOISE FIGURE FOR 200Ω SOURCE.
5- INDICATOR TIME CONSTANT - ½ SECOND
MAXIMUM ERROR IN MEASUREMENT IN DB

RANGE OF INSERTION LOSS MEASUREMENT IN DB AS LIMITED BY NON-LINEARITY OF R.F. SQUARE LAW DETECTOR AND NOISE FLUCTUATIONS OF AUDIO SYNCHRONOUS DETECTOR.

Figure 6

Acknowledgements

This work was supported by the Wright Air Development Center and Air Materiel Command under the direction of Messrs. E. Birochak and R. W. Stolzenbach.

APPENDIX

A. Representation of Signal and Noise

If we have random noise voltage on a particular circuit and at a particular time make an instantaneous measurement of this voltage, we cannot, of course, say what this measurement will give, but we can indicate the probability of this measurement's having certain values. The probability of the different values of this voltage can be expressed by:

$$P(v)dv = \frac{1}{\sqrt{2\pi}\,\sigma}\, e^{-\frac{v^2}{2\sigma^2}}\, dv \qquad (1)$$

This is the Gaussian or one dimensional normal distribution. P (v) represents the probability of a particular value of v occurring where v is the voltage. σ^2 is the variance, and actually corresponds to the average square value of the voltage, averaged over all time. In other words, the root mean square value of the voltage would be given by σ. The average value of the voltage is clearly zero, since v occurs only as the square in Equation (1), and positive and negative values of v are equally probable. In the case of thermal noise, with which we are principally concerned,

$$\overline{E^2} = \sigma^2 = 4RkT\Delta F \qquad (2)$$

where E is the RMS value of the thermal noise voltage,
R is the resistive component of the impedance in ohms,
T is the absolute temperature,
k is the boltzmann constant and
Δ F is the bandwidth.

If, instead of analyzing the voltage by a number of instantaneous samplings of its value, we pass the signal through a series of sharply tuned filters, we will find that the signal represents the summation of many individual frequency components and can be expressed by:

$$v = \int_{w_1}^{w_2} V(w)\cos[wt + \phi(w)]\, dw \qquad (3)$$

Here V (w) is the amplitude of the individual frequency component and ϕ (w) represents its phase with respect to some arbitrary point in time. It can be shown that V (w) is also a statistical variable; i.e., its value is not completely determined, but only the probability of its having particular values can be expressed. For thermal noise the probability of a particular value of V (w) can be expressed by a two dimensional normal distribution.

$$P(V)dV = \frac{V}{\sigma^2}\, e^{\frac{-V^2}{2\sigma^2}}\, dV \qquad (4)$$

Notice that in this case the presence of the variable outside the exponential indicates that the average value of V is no longer zero. That is, there is some most likely value of V other than zero and some fluctuation about this value.

If we wish to consider this noise in conjunction with the signal of frequency $\Omega/2\pi$, it is well to change the form of Equation (3) as follows:

$$v = \int_{w_1}^{w_2} V(w)\cos[(w-\Omega)t + \Omega t + \phi(w)]\, dw \qquad (5)$$

If we now use a formula for the cosine of the sum of two angles, we obtain:

$$v = \int_{w_1}^{w_2} V(w)\cos(\Omega t)\cos[(w-\Omega)t + \phi(w)] - V(w)\sin(\Omega t)\sin[(w-\Omega)t + \phi(w)]\, dw \qquad (6)$$

This can be recognized as an expression for a carrier of frequency $\Omega/2\pi$ modulated at a frequency $(w-\Omega)/2\pi$, where $\phi(w)$ represents the phase of the modulation. In fact, if we change variables, writing W for $w-\Omega$, we have:

$$v = \int_{w_1}^{w_2} V(W)\left\{\cos(\Omega t)\cos[Wt + \phi(w)] - \sin(\Omega t)\sin[Wt + \phi(W)]\right\} dW \qquad (7)$$

The limits on the summation have been changed in accordance with the new variable, W, having been substituted for w. If we rewrite the expression in slightly different form, we have:

$$v = \cos(\Omega t)\int_{w=-B/2}^{B/2} V(W)\cos[wt + \phi(W)]\,dW - \sin(\Omega t)\int_{w=-B/2}^{B/2} V(W)\sin[wt + \phi(W)]\,dW \qquad (8)$$

The expression is now written so as to bring out its form as a modulated signal. The limits of the summation have been changed to $W = \pm B/2$ to indicate the fact that this noise signal has been passed through an amplifier with a band pass of B centered at a frequency of Ω

B. Linear Or Envelope Detectors

In order to find the envelope of this signal, which is essentially what an envelope detector would do, we must first make the assumption that the bandwidth of the I.F. is narrow compared to the carrier frequency. If this condition does not hold, the envelope of the signal has no significance. Assuming that the band width of the I.F. is sufficiently narrow, the envelope of the signal is given by the square root of the sum of the squares of the amplitudes of the sine and cosine terms. These amplitudes are represented by the integrals. Therefore, in order to find the envelope, one must find the square root of the sum of the squares of the two integrals in Equation (8). This appears to be a task of some complexity which cannot be attacked directly. However, the similarity of these integrals to that of Equation (3) is apparent and except for the limited band width, these integrals represent white noise similar to Equation (3). That is, if we were to measure the instantaneous value of the amplitude of the Cos Ωt in Equation (8), the probability of obtaining particular values for the amplitude would be expressed by an equation similar to Equation (1); i.e., the amplitude of this sinusoid expressed by the integral in Equation (8) is a function of time which can also be represented by a normal distribution of the type of Equation (1). These considerations also hold for the amplitude of the term in $\sin \Omega t$ in Equation (8). If we combine these two terms into a single sinusoid, its amplitude will be the sum of the squares of the amplitudes of the cos and sin terms; i.e., the sum of the squares of two distributions of the type of Equation (1). It can be shown that the sum of the squares of two distributions of the type of Equation (1) is a two dimensional normal distribution of the type illustrated in Equation (4).

The amplitude of the combined expression can, therefore, be given by:

$$P(R)\,dR = \frac{R}{\sigma^2}\, e^{\frac{-R^2}{2\sigma^2}}\, dR \qquad (9)$$

where R is the amplitude of the signal at frequency, $\Omega/2\pi$ and P (R) is the probability of occurrence of a particular value of R. Here again, as with our previous two dimensional normal distribution, the average value of R is not zero. It can be obtained according to its definition as follows:

$$\bar{R} = \frac{\int_0^\infty R P(R)\,dR}{\int_{-\infty}^{\infty} P(R)\,dR} = \int_0^\infty \frac{R^2}{\sigma^2}\, e^{\frac{R^2}{2\sigma^2}}\, dR \qquad (10)$$

$$= \sigma\sqrt{\frac{\pi}{2}} \cong 1.25\sigma$$

In this derivation represented by Equation (10), it is recognized that the integral of P (R) dR from minus infinity to infinity is one, and the integration of R^2 P (R) dR is carried out by parts. In order to examine the significance of the numerical factor $\sqrt{\pi/2}$ in Equation (10), let us suppose that our signal, instead of being modulated by a stochastic signal, i.e., a signal which can be expressed only in terms of probability, is modulated by a sine wave. That is, that our signal actually consists of two high frequency signals 90 degrees out of phase, each modulated by a simple sinusoid. Let us assume that the RMS value of each of the sinusoids is σ, we will then find that our expression corresponds to that of Equation (10), i.e., the average value of the envelope will vary between $.9\sigma$ and $1.414\,\sigma$, depending upon the phase relationship between the two modulating signals.

In order to investigate the phenomenon of signal suppression, let us introduce a signal along with the noise. If we return now for the moment to Equation (8) and introduce a signal of the form S cos Ω t, Equation (8) will be transformed into:

$$v = \cos(\Omega t)\left\{S + \int_{w=-B/2}^{B/2} V(W)\cos[wt + \phi(W)]\,dW\right\} - \sin(\Omega t)\int_{w=-B/2}^{B/2} V(W)\sin[wt + \phi(W)]\,dW \qquad (11)$$

We have merely added the signal. If we now proceed as we did with the noise alone, recognizing that the two integrals in Equation (11) can be represented by normal distributions, we can find the distribution which represents the sum of the squares of the amplitudes of the two quadrature components cos Ωt and sin Ωt in Equation (11). This result will be analogous to what we found in Equation (9). However, in this case it will be:

$$P(R)dR = \frac{1}{\sigma^2} R e^{-\frac{R^2+S^2}{2\sigma^2}} I_o\left(\frac{RS}{\sigma^2}\right) dR \quad (12)$$

where I_o is the modified bessel function of the first kind of zero order. It seems worthwhile to digress momentarily at this point to clarify the mystery of the sudden appearance of the bessel function. In transforming from the two normal distributions to the distribution in R, the process is essentially one of considering the variables of the two normal distributions as rectangular coordinates values and transforming to polar coordinates. In the transformation to polar coordinates, the integral

$$\int_0^{2\pi} e^{-\frac{R\sin\Theta}{\sigma^2}} d\Theta$$

occurs and the value of this integral is simply I_o, the modified bessel function.

Equation (12) expresses the probability of occurrence of a particular value of the envelope of the signal and noise as a function of the value, the signal and the noise. R is the value of the envelope, S is the amplitude of the signal and σ is the RMS value of the noise. An envelope detector, as used in practice, would consist of some circuit which puts out a D.C. voltage, proportional to the value of the envelope, and some smoothing circuit following this which essentially averages this D.C. voltage and thus finds the average value of the envelope. Therefore, what we wish to find is the average value of R. This will correspond to the output of our smoothing circuit. We will consider averaging over large periods of time. This average value of the envelope is represented by:

$$\bar{R} = \int_0^\infty R P(R) dR = \int_0^\infty \frac{R^2}{\sigma^2} e^{-\frac{R^2+S^2}{2\sigma^2}} I_o \left(\frac{RS}{\sigma^2}\right) dR = \frac{\sigma}{\sqrt{2\pi}} e^{-\frac{S^2}{4\sigma^2}} \left[\left(1+\frac{S^2}{2\sigma^2}\right) I_o\left(\frac{S^2}{4\sigma^2}\right) + \frac{S^2}{2\sigma^2} I_1\left(\frac{S^2}{4\sigma^2}\right)\right] \quad (13)$$

where I_1 is the modified bessel function of the first kind and first order. The integration in this case is fairly complex.

To illustrate the mechanism of signal suppression by noise, we will examine equation (13) in the case of limiting values of signals and noise; i.e., where S/σ is large and where S/σ is small. Consider first the case where $S/\sigma \gg 1$

$$I_o\left(\frac{S^2}{4\sigma^2}\right) = \frac{\sigma}{\sqrt{\pi} S}\left[1+\frac{\sigma^2}{2S^2}\right] e^{\frac{S^2}{4\sigma^2}}$$

$$I_1\left(\frac{S^2}{4\sigma^2}\right) = \frac{\sigma}{\sqrt{\pi} S}\left[1-\frac{3\sigma^2}{2S^2}\right] e^{\frac{S^2}{4\sigma^2}} \quad (14)$$

$$\bar{R} = \frac{S}{\pi\sqrt{2}} + \frac{\sigma^2}{2\pi S\sqrt{2}} + \text{higher terms in } \frac{\sigma}{S}$$

The two principal terms which constitute the average value of R are listed in Equation (14). It is apparent that for the case under consideration the first term is dominant and since this is a term which varies linearly with S, the output of our envelope detector will be directly proportional to the signal. Note that σ does not occur in this term, so that essentially the output of the detector is independent of the noise.

Let us consider now the opposite extreme where the signal is considerably less than the noise, here $S/\sigma \ll 1$

$$I_o\left(\frac{S^2}{4\sigma^2}\right) = 1 + \frac{S^4}{16\sigma^4} + \cdots\cdots$$

$$I_1\left(\frac{S^2}{4\sigma^2}\right) = \frac{S^2}{2\sigma^2}\left(1+\frac{S^2}{32\sigma^2}\right) + \cdots\cdots \quad (15)$$

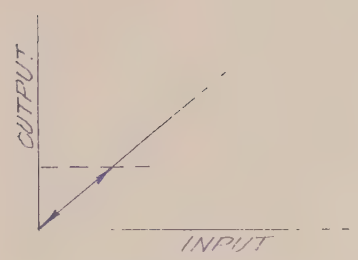
$$e^{\frac{-S^2}{4\sigma^2}} = 1 - \frac{S^2}{4\sigma^2} + \cdots\cdots$$

$$\bar{R} = \frac{\sigma}{\sqrt{2\pi}} + \frac{S^2}{4\sigma\sqrt{2\pi}}$$

In this case we note that the average value of the envelope no longer depends linearly on the signal. The dependence appears to be square law with the value of S appearing in the denominator of the term, so that, in effect, for a constant value of noise, the average output of our detector when a signal is applied will increase by an amount pro-

portional to the square of that signal. It will also be inversely proportional to the first power of the noise; i.e., the larger the noise , the smaller will be the change in output of the detector when the signal is applied. This is the phenomenon which constitutes signal suppression by noise. It is similar to a quadrature combination of the signal and noise; i.e., if we consider the square root of the sum of the squares of the signal and noise amplitudes we will find that this square root has a very similar dependence on the relative magnitudes of the signal and noise. For large signal the square root is relatively independent of the noise and depends directly upon the signal amplitude in a linear way; whereas, for large noise the converse is true, the square root will depend principally upon the magnitude of the noise in a linear way and only very slightly upon the magnitude of the signal amplitude.

In order to obtain some physical understanding of the phenomenon of signal suppression by noise in linear detectors, consider Figure 1. This represents the curve of output vs. input for a linear detector. If the amplitude of the signal is as indicated by the arrow, the average output of the detector will be somewhere slightly below the dotted line. Consider now biasing the detector with an amount of direct current which is greater than the amplitude of the signal. We now find that the signal operates on a portion of the linear detector curve away from zero completely, so that the average value of the output is completely unaffected by the presence of the signal. That is, so long as the signal and bias are such that the signal does not carry the detector through its zero point, it will on the average add as much to the bias as it subtracts and, therefore, contributes nothing to the output. The effect of noise is very similar to a bias in that when the noise is larger than the signal, during an appreciable portion of the time the signal is in the condition that it does not carry the detector through the zero point and, consequently, for this appreciable portion of the time, the signal does not contribute to the average output of the detector. The larger the noise, the smaller the percentage of time during which the signal is able to carry the detector through the zero point, and thus make a contribution.



Linear Detector
Figure 1

C. Square Law Detectors

Let us make clear at the outset what is meant by a square law detector. It is necessary to differentiate between the square of the output of a linear detector and the output of a true square detector. If we square the output of a linear detector, we have effectively the square of the average linear value. On the other hand, a true square law detector gives the average square of the signal. We are discussing here a detector whose average value is proportional to the average value of the square of the signal. This is essentially the way in which bolometers perform. The input power is proportional in this case to the square of the signal and the output is proportional to the average value of the input power, so that the output is, in fact, proportional to the average square of the input.

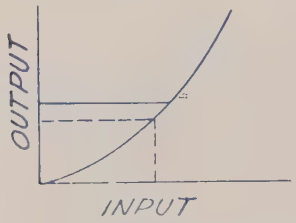
Our representation of the combination of signal and noise, which we made in Equation (11) is still quite adequate for the present problem and the probability distribution of the amplitude is given, as before, by Equation (12). However, we are now interested in the average value of the square of the amplitude and in place of Equation (13), we should use

$$\overline{R^2} = \int_0^\infty R^2 P(R)\,dR = \int_0^\infty \frac{R^3}{\sigma^2} e^{-\frac{R^2+S^2}{2\sigma^2}} I_0\left(\frac{RS}{\sigma^2}\right) dR = \tfrac{1}{2}S^2 + \sigma^2 \qquad (16)$$

In this case the definite integral gives a very simple result, namely, the mean square value of the signal plus the mean square value of the noise. It should be noted that whereas σ is the root mean square value of the noise, S is the amplitude of the signal and not

its root mean square value. This accounts for the factor 1/2. Thus we see that in the case of a bolometer the addition of a signal will always give the same change in the average value of the output, independent of the amount of noise present; i.e., there is no suppression of signal by noise in this case.

In order to understand simply why this should be, consider Figure 2. If the dotted lines represent the input and the output of the detector, with no signal present, the effect of the signal is to change the input above and below this average value by the same amount. However, due to the curvature of the characteristic, the output will not be changed an equal amount above and below its quiescent value. The increase in output when the signal is positive will be greater than the decrease when the signal is negative and, consequently, the average value of the output will be increased by the signal. If the detector is a perfect square law, this increase in the average value of the output will be the same, regardless of the bias point of the detector. Consequently, the considerations relating to Figure 2 do not hold here and one would expect that the contribution of the signal to the average output will be independent of the level of the noise. Consequently, the result of Equation (16) is not too surprising.



Square Law Detector
Figure 2

There is one more point of view from which square law detection can be examined and it seems worthwhile to do this, since it will serve in a small way as an introduction to the problem of synchronous detection. If we disregard for the moment the probability distributions and examine Equation (11) with a view to the performance of the square law detector, we recognize that the output of the detector will be proportional to the average square of v. The square of v will be an extremely complicated expression; however, since we are only concerned with the average of this, we can delete from this expression all those terms whose time average is zero. The time average over a long period of the product of two sinusoids of different frequencies is zero, so that we can eliminate these from further consideration. Further, the time average over a long period of the product of the sine and cosine terms is zero, even though these two terms may be of same frequency. Also, the time average of $\sin^2$ and $\cos^2$ = 1/2. These facts lead us to:

$$\overline{v^2} = \frac{S^2}{2} + \tfrac{1}{2}\int_{-B/2}^{B/2} V^2(\omega) + \tfrac{1}{2}\int_{-B/2}^{B/2} V^2(\omega) + \tfrac{1}{2}\int_{-B/2}^{B/2} V^2(\omega) = \frac{S^2}{2} + \int_{-B/2}^{B/2} V^2(\omega) = \frac{S^2}{2} + \sigma^2$$

where it is necessary to recognize that the integral of the amplitude, $(V(w))^2$ actually represents the mean square value of the noise.

D. Synchronous Detector

The action of a synchronous detector is essentially the multiplication of the combination of signal and noise, expressed by Equation (11), by a sinusoidal signal of frequency $\Omega/2\pi$ and the subsequent averaging of this product. If the synchronous signal(a) by which this combination of signal and noise is multiplied is given by:

$$a = A\cos(\Omega t) \qquad (18)$$

the process of multiplication and avereging can be expressed mathematically by:

$$P = \frac{1}{T}\int_{t}^{t+T} v \cdot a\, dt \qquad (19)$$

where T is of the order of the time constant of the lowpass filter which follows the detector. Here T is very much greater than $\frac{2\pi}{\Omega}$, the period of the fundamental frequency.

In order that Equation (11) will have significance as expressing a modulated signal, it is necessary to assume, as before, that the band width B is small compared to the carrier frequency Ω. With this in mind, an examination of Equation (11) shows that the $\cos\Omega t$ and $\sin\Omega t$ are rapidly varying quantities. All of the quantities under the integral signs are slowly varying. Therefore, in

order to simplify the mathematics, we can consider our averaging in two parts. First, we can average over a cycle of Ω, the rapidly varying part, and then consider the subsequent averaging of these average values over the long time period T. We note that in the product of Equation (11) by Equation (18) the sin Ω t term in Equation (11) will yield a term of the form of the product of:

$$\sin \Omega t \times \cos \Omega t$$

which has the average value of over one cycle of zero. Thus sin Ω t can, therefore, be dropped, as it will not contribute anything to the average value expressed by Equation (19). The product of the two cos Ω t terms will give an average value over each cycle of 1/2. Our output P, expressed by Equation (19), is, therefore, reduced to:

$$P = \frac{aS}{2} + \frac{a}{2T}\int_{W=-B/2}^{W=+B/2}\int_{t}^{t+T} V(W)\cos \left[Wt + \phi(W)\right] dt\, dW \qquad (20)$$

The integration over T can easily be performed, giving:

$$P = \frac{aS}{2} + \frac{a}{2T}\int_{W=-B/2}^{W=+B/2} \frac{V(W)}{W}\left\{\sin\left[W(t+T) + \phi(W)\right] - \sin\left[Wt+\phi(W)\right]\right\} dW \qquad (21)$$

Some understanding of the action of synchronous detectors can be obtained by examining Equation (21) in detail. Note that the difference in the two sine terms can be positive or negative. For large values of wt, the two sine terms are substantially independent of each other and each may independently have values ranging from -1 to + 1; thus the maximum value of the integrand is $\frac{2V(W)}{W}$. The minimum value is $\frac{-2V(W)}{W}$ and the average value is zero. Thus clearly the integral is as likely to be positive as negative and its average value will be zero. However, an individual integral will have a statistical fluctuation about this average value. The presence of the w in the denominator of the integrand indicates that for large values of w, a very small contribution to the output will be made. Since w, is the difference between the signal frequency and the frequency of the noise, this means that noise components at frequencies considerably different from that of the signal will have no appreciable effect upon the output T. It might appear at first glance that since the integral is carried through the region where w goes to zero the integrand will at this point be infinite. However, for small values of w the difference between the two sine terms must of necessity approach zero, so that when w is zero the integrand is $\frac{0}{0}$, which is indeterminate.

The synchronous detector, like the square law detector, exhibits no signal suppression by noise; further, the average contribution of the noise to the detector output is zero. Comparing Equations (16) and (21), we see that the term $S^2/2$ in Equation (16) for the square law detector is replaced by $aS/2$, in the case of the synchronous detector, where (a) is the large synchronizing signal.

APPENDIX 2

Fluctuations of Detector Output Due to Noise in Synchronous Detector.

The problem posed here is as follows: A signal plus noise is amplified and synchronously detected. The average output of the detector depends, of course, only on the signal amplitude, but the noise will cause fluctuations of the output about its average value. We are interested in the magnitude of these fluctuations as a function of:

- Ba - The bandwidth of the amplifier
- Bf - The bandwidth of the filter following the detector
- S - The signal amplitude going into the detector
- N - The white noise amplitude going into the detector

The instantaneous voltage at the input to the detector can be described by:

$$P(v)dv = \frac{1}{\sigma\sqrt{2\pi}} e^{-\frac{v^2}{2\sigma^2}} dv \qquad (1)$$

where P (v) is the probability of occurrence of a particular value of voltage v.

σ is the root mean square of the voltage. This significance of σ can be established by finding the rms value of v as:

$$\overline{v^2} = \frac{\int_{-\infty}^{+\infty} v^2 P(v) dv}{\int_{-\infty}^{+\infty} P(v) dv} = \frac{\int_{-\infty}^{+\infty} \frac{v^2}{\sigma\sqrt{2\pi}} e^{-\frac{v^2}{2\sigma^2}} dv}{1} = \sigma^2 \qquad (2)$$

The integration is performed using a table of definite integrals.

If we view the voltage in terms of its frequency components, instead of in terms of its probable instantaneous values, we have:

$$v = \int_0^{Ba} a \sin(2\pi f t + s)\, df \qquad (3)$$

where a and s are random functions of f. We can also express the mean square value of v in terms of the frequency components:

$$\overline{v^2} = \lim_{t\to\infty} \frac{1}{t}\int_0^t v^2 dt = \lim_{t\to\infty} \frac{1}{t}\int_0^t \left[\int_0^{Ba} a \sin(2\pi f t + s)\, df\right]^2 dt \qquad (4)$$

We cannot evaluate Equation (4) directly, but since $\overline{v^2} = \sigma^2$ from Equation (2), it follows that:

$$\lim_{t\to\infty} \frac{1}{t}\int_0^t \left[\int_0^{Ba} a \sin(2\pi f t + s)\, df\right]^2 dt = \sigma^2 \qquad (5)$$

Since a synchronous detector operates by forming the product of the input with the synchronizing signal, the synchronous detector output V, before filtering, is given by:

$$V = v \cdot v_S = v \cdot A \sin(2\pi f_o t) \qquad (6)$$

where $v_S = A \sin(2\pi f_o t)$ is the synchronizing signal.

$$V = A\int_0^{Ba} a \sin 2\pi f_o t \sin(2\pi f t + s)\, df$$

$$= \frac{A}{2}\int_0^{Ba} a \cos\left[2\pi(f_o - f)t - s\right] - a\cos\left[2\pi(f_o + f)t + s\right] df \qquad (7)$$

Let us suppose that $(f_o + f) > Bf$; that is, we will assume that only the difference frequency is of interest, the sum of the two frequencies being higher than our output filter band pass. Then:

$$V = A/2\int_0^{Ba} a \cos\left[2\pi(f_o - f)t - s\right] df$$

let $(f_o - f) = fn$
then $df = -dfn$
and the new limits of integration are fo to fo-Ba
Then:

$$V = -A/2\int_{f_o}^{f_o - Ba} a \cos(2\pi f n t - s)\, df n \qquad (8)$$

This equation is very similar to Equation (3). The integrands are statistically the same and the band width of integration is also the same. Therefore, except for the factor A/2, the voltage defined by Equation (8) must have the same mean square value as that defined by Equation (3). Thus:

$$\overline{V^2} = \frac{A^2}{4}\sigma^2 \qquad (9)$$

Now suppose that the output of the synchronous detector is filtered, so that only signals of frequency less than Bf can pass:

$$V = -A\int_{Bf}^{-Bf} a\cos(2\pi f_n t - S)\,df_n \qquad (10)$$

Since the various frequency components are completely uncorrelated, they add in an r.m.s. fashion, and the mean square value of the integral must be proportional to the range of integration (2Bf). Therefore:

$$\overline{V^2} = \frac{A^2}{4}\sigma^2\,\frac{2Bf}{Ba} \qquad (11)$$

and the r.m.s. value of the fluctuation is:

$$\sqrt{\overline{V^2}} = A\sigma\sqrt{\frac{Bf}{2Ba}} \qquad (12)$$

Since the r.m.s. noise input is

$$\sqrt{\overline{v^2}} = \sigma \qquad (13)$$

$$\frac{\text{r.m.s. output fluctuation}}{\text{r.m.s. noise input}} = A\sqrt{\frac{Bf}{2Ba}} \qquad (14)$$

A is the ratio of D.C. signal out of the synchronous detector to synchronous signal in (r.m.s. value). This can readily be shown by letting v in Equation (6) be a signal of synchronous frequency. Therefore:

$$\frac{\text{r.m.s. output fluctuation}}{\text{r.m.s. noise input}} = \frac{\text{signal output}}{\text{signal input}}\sqrt{\frac{Bf}{2Ba}}$$

or

$$\frac{\text{signal to noise in}}{\text{signal to noise out}} = \sqrt{\frac{Bf}{2Ba}} \qquad (15)$$

where it is understood that "Signal to Noise Out" means the ratio of the r.m.s. output fluctuation to the average D.C. output. Thus, if the signal to noise ratio desired in the output is 100 (fluctuations generally 1% of reading, sometimes 2%, and rarely 3%),

$$Bf = 2Ba \times \left(\frac{S/n}{100}\right)^2 \qquad (16)$$

where Bf equivalent band width of the synchronous detector output filter
Ba band width of amplifier preceding synchronous detecor
S/N Signal to noise voltage ratio measured at amplifier output
100 represents desired S/N or D.C. reading to fluctuation ratio at the synchronous detector output

For example, if S/N = 1, Ba = 5000

$$Bf = 5000 \times \frac{2}{10{,}000} = 1 \text{ cycle/sec.}$$

which is a time constant of .5 seconds.

Formula 15 forms the basis for the data plotted in the left part of Figure 5 and the center portion of Figure 6 of the main paper.

B. Fluctuations of Detector Output Due to Noise in Square Law Detector

The problem posed here is as follows: White noise passes through an amplifier and is detected by a square law detector. The detector output will have a certain average value and a certain fluctuation about that average. We wish to find the root mean square fluctuation of the detector output from its average value as a function of:

Ba - The band width of the amplifier
Bf - The bandwidth of the filter following the detector
S - The sensitivity of the detector in

$$\frac{\text{Output volts}}{(\text{Input volt})^2}$$

We shall assume that the instantaneous value of the voltage at the input to the detector has a one dimensional normal distribution; i.e.,

$$P(v) = \frac{1}{\sigma\sqrt{2\pi}}\,e^{-\frac{v^2}{2\sigma^2}} \qquad (17)$$

The detector output V is given by:

$$V = Sv^2$$

and $P(V) = P(Sv^2)$. That is, the probability of a given value of detector output voltage is the same as the probability of the corresponding value of detector input voltage.

The average value of v^2 is given by:

$$\overline{v^2} = \frac{1}{\sigma\sqrt{2\pi}}\int_{-\infty}^{\infty} v^2 e^{\frac{-v^2}{2\sigma^2}} dv =$$

$$= \frac{2}{\sigma\sqrt{2\pi}}\int_{0}^{\infty} v^2 e^{\frac{-v^2}{2\sigma^2}} dv \qquad (18)$$

from Dwight's tables of integrals, page 494

$$\overline{v^2} = \frac{2}{\sigma\sqrt{2\pi}}\sqrt{2\pi}\,\sigma \cdot \frac{2\sigma^2}{4} = \sigma^2$$

Therefore:

$$\overline{V} = S\,\overline{v^2} = S\sigma^2 \qquad (19)$$

$$\overline{(V-\overline{V})^2} = \overline{(Sv^2 - S\sigma^2)^2} = S^2\overline{(v^2-\sigma^2)^2} \qquad (20)$$

where $\overline{(V-\overline{V})^2}$ is the mean square deviation of V from the mean value

$$\overline{(v^2-\sigma^2)^2} = \frac{1}{\sigma\sqrt{2\pi}}\int_{-\infty}^{\infty}(v^2-\sigma^2)^2 e^{\frac{-v^2}{2\sigma^2}} dv$$

$$= \frac{2}{\sigma\sqrt{2\pi}}\int_{0}^{\infty}(v^4 - 2\sigma^2 v^2 + \sigma^4) e^{\frac{-v^2}{2\sigma^2}} dv \qquad (21)$$

using Dwight's tables again, we have:

$$\overline{(v^2-\sigma^2)^2} = \frac{2}{\sigma\sqrt{2\pi}}\left[\frac{3}{8}(2\sigma^2)^2\sqrt{2\pi}\,\sigma - 2\sigma^2\,\frac{1}{4}2\sigma^2\sqrt{2\pi}\,\sigma + \sigma^4\frac{\sqrt{2}}{2}\sigma\sqrt{\pi}\right]$$

$$= 2\left[\frac{3}{2}\sigma^4 - \sigma^4 + \frac{\sigma^4}{2}\right] = 2\sigma^4$$

Therefore: $2S^2\sigma^4$ = mean square deviation of V from its means value $\therefore$ r.m.s. deviation of V from its meen value

$$\sqrt{2}\,S\sigma^2 = \sqrt{2}\,\overline{V} \qquad (23)$$

At first glance, it may appear contradictory that the r.m.s. fluctuation in V is greater than its average value, since V cannot be negative. However, V has a skew distribution. It is below its average value most of the time. Its excursions above its average value are infrequent, but large. Thus, from equation (1):

$$P(v)dv = \frac{1}{\sigma\sqrt{2\pi}}\, e^{-\frac{v^2}{2\sigma^2}} dv$$

$$V = Sv^2 \qquad (24)$$

$$dV = 2Sv\,dv = 2\sqrt{SV}\,dv$$

Hence, the probability distribution of V is given by:

$$P(V) = \frac{1}{\sigma\sqrt{2\pi SV}}\, e^{-\frac{V}{2S\sigma^2}} dV \qquad (25)$$

where a factor of 2 has been introduced to account for the fact that a particular value of V corresponds to a positive or negative value of v.

Figure 1 is the plot of the distribution given by Equation (25). It illustrates that while the meen value of V is $S\sigma^2$, the most probable value is zero, if the output is unfiltered. A qualitative sketch of the distribution with a filter following the detector is also included in Figure 1.

ENGINEERING NOTES NO. 2
March 1955

INSERTION LOSS TEST SETS USING SQUARE LAW DETECTORS

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Summary

This paper discusses theory and errors, when measuring radio frequency attenuation, using imperfect square law detectors.

The audio modulation of a radio frequency source is detected by the resistance modulation of a bolometer producing an audio frequency voltage.

For ideal square law detectors, the variation of the audio frequency level in db, using a constant current source, is twice the db change of the radio frequency level.

A theoretical analysis of the error due to deviation from square law is given. Experimental data regarding this error based on the static characteristic and independent experimental data relying upon differential R.F. measurements are furnished and are in agreement with the theory.

Applying this error analysis to typical barretters, accuracies of .01 db are feasible for a measurement of a 20 db loss.

A simple test set which has been reduced to practice is discussed. It permits R.F. loss measurements of 30 db with an accuracy of .1 db.

Introduction

This paper was presented during the Radio Conference on September 8, 1954 at the Boulder Laboratories of the National Bureau of Standards. The program was supported by the Communication and Navigation Laboratory of the Wright Air Development Center under the direction of Messrs. E. Birochak, Project Engineer, and R. W. Stolzenbach, Chief of the Calibration and Standards Section.

A theory is demonstrated in the first part which shows that insertion loss measurements can be made to high accuracies using barretters. The limitations in reducing this to practice are manifold. However, theoretically it is possible to measure losses up to 10 db with an accuracy of .01 db.

Since the square-law range of barretters is limited on the high power side, it was of interest to explore the limitations at the low power side due to noise. The efficiency of detectors in the presence of noise was looked into and relationships showing the fluctuations of a coherent detector output due to noise were derived. As a result of these studies, a coherent detector operating at 1000 cps. was constructed which extended the useful range over an average detector by 40 db audio or 20 db RF without increase of RF power.

Several good papers are available on the non-linearity of linear mixers.[1,2,3,4,5] It is well-known that its non-linearity depends upon the ratio of maximum signal voltage to local oscillator voltage and its static characteristic.

As is shown below, the corresponding magnitudes in a square law detector are the ratio of maximum RF power to bias power and the static characteristic of the barretter.

Barretters commercially available, such as, e.g., the P.R.D. #610-A, work rather well at 400 cps. Its response at 400 cps is about 2 db below the one obtained by D.C. keying and at 1000 cps, it is only 4 db below the D.C. response. Operation at these audio frequencies reduces the influence of microphonics and magnetic coupling from A.C. power supplies. The barretters proved to be very stable and accurate elements useful for precise measurements.

The theory was reduced to practice in the frequency range 50 to 10,000 mc using coaxial construction of 50 ohm characteristic impedance and in waveguide at X-Band for insertion loss measurements of fixed and variable attenuators.

Attenuation Measurement

For measuring attenuation, one can use various methods, such as, e.g.,

1. The measurement of input and output voltage.
2. Comparing the unknown with an R.F. standard attenuator by series substitution in a single channel method.

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3. Comparing the unknown with a standard attenuator by parallel balance in a two channel method.

4. The substitution may also be made at an I.F. frequency resulting in the well-known heterodyne insertion loss measurement.

Methods 2 and 3 depend upon the accuracy of an attenuation standard, usually a waveguide below cut-off attenuator.

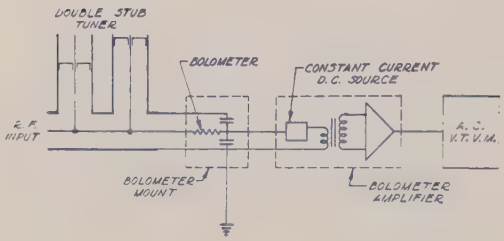
Method 4 relies on the linearity of a heterodyne mixer and the accuracy of the I.F. attenuator.

Methods 2 through 4 are linear methods since for every db of the unknown attenuator, the same amount of decibels is removed from the standard in single channel methods or is added in two channel methods.

5. The "Ballantine Voltmeter" method, using a square law detector.[6,7,8]

Theory of Non-Linearity

Figure 1 shows the typical use of a bolometer for detecting an R.F. carrier which is square wave modulated at an audio frequency.



USE OF BOLOMETER AS SQUARE LAW DETECTOR

Figure 1

Precautions

When making insertion loss measurements, certain standard precautions must be observed, such as, e.g.,

1) Use of a well matched load.
2) Use of a well matched source.
3) Frequency and amplitude stability of the source.

4) Operation of the detector at optimum bias point to limit its non-linearity.

5) Limitation of the maximum signal to limit the non-linearity of the detector.

6) Low crosstalk.

7) In case of two-channel methods, sufficient r.f. isolation to avoid inaccuracies due to a second signal path.

Theoretical Derivation of Error Due To Deviation from Square Law

It has been found experimentally that Wollaston type barretters which are cooled mainly by convection follow an exponential law. In order to compute the error in insertion loss measurement, it is more convenient to express the bolometer resistance in exponential form.

$$R = R_o + J P^n \qquad (1)$$

where R is the barretter resistance

- R_o is the cold barretter resistance
- P is the power dissipation (D.C. R.F.)
- ΔP = R.F. power to barretter with attenuator out
- $\Delta P'$ = R.F. power to barretter with attenuator in
- J is a constant depending on the individual barretter
- n is approximately 0.9 for many commercially available barretters.

If $n = 1$, the barretter would be a perfect square law detector. Let us consider that in measuring an attenuator by the method outlined in the preceding paragraph, the assumption is arbitrarily made that the barretter is a perfect square law detector, i.e. that:

$$X = 10 \cdot \log \frac{\Delta P}{\Delta P'} = 10 \cdot \log \frac{\Delta R}{\Delta R'} \, DB \qquad (2)$$

Using Equation (1), the difference in power ratios as indicated by a barretter for $n \neq 1$ and a perfect square law detector is computed below.

$$\Delta R = R_o + J(P_1 + \Delta P)^n - (R_o + J P_1^n) \qquad (3)$$
$$= J\left[(P_1 + \Delta P)^n - P_1^n\right]$$

where P_1 is the D.C. bias power

Similarly

$$\Delta R' = J\left[(P_1 + \Delta P')^n - P_1^n\right] \qquad (4)$$

$$\frac{\Delta R}{\Delta R'} = \frac{(P_1+\Delta P)^n - P_1^n}{(P_1+\Delta P')^n - P_1^n} \qquad (5)$$

If we now expand $(P_1 + \Delta P)^n$ by means of the binomial formula, retaining the first three terms, due to good convergence

$$(P_1+\Delta P)^n = P_1^n + nP_1^{n-1}\Delta P + \frac{n(n-1)}{2} P_1^{n-2}(\Delta P)^2 \qquad (6)$$

and similarly with $(P_1 + \Delta P)^n$. We find:

$$\frac{\Delta R}{\Delta R'} = \frac{2nP_1^{n-1}\Delta P + n(n-1)P_1^{n-2}(\Delta P)^2}{2nP_1^{n-1}\Delta P' + n(n-1)P_1^{n-2}(\Delta P')^2} =$$

$$= \frac{\Delta P}{\Delta P'} \cdot \frac{2P_1 + (n-1)\Delta P}{2P_1 + (n-1)\Delta P'} = \qquad (7)$$

$$= \frac{\Delta P}{\Delta P'}\left\{1 + \frac{(n-1)(\Delta P - \Delta P')}{2P_1 + (n-1)\Delta P'}\right\}$$

We can now neglect $(n-1)\Delta P'$ in the denominator as compared to $2P_1$, so that to a sufficient degree of approximation,

$$\frac{\Delta R}{\Delta R'} = \frac{\Delta P}{\Delta P'} + \frac{(n-1)\Delta P(\Delta P - \Delta P')}{2P_1 \Delta P'} \qquad (8)$$

so that if the barretter is arbitrarily assumed to be a perfect square law detector, the attenuator will be measured as

$$X_m = 10 \cdot \log \frac{\Delta R}{\Delta R'} =$$

$$= 10 \cdot \log\left[\frac{\Delta P}{\Delta P'} + \frac{(n-1)\Delta P(\Delta P - \Delta P')}{2P_1 \Delta P'}\right] db \qquad (9)$$

while its actual value is:

$$X = 10 \cdot \log \frac{\Delta P}{\Delta P'}$$

The error in db is, therefore,

$$X_m - X = 10\left\{-\log \frac{\Delta P}{\Delta P'} + \log\left[\frac{\Delta P}{\Delta P'} + \frac{(n-1)\Delta P(\Delta P - \Delta P')}{2P_1 \Delta P}\right]\right\} \qquad (10)$$

$$= 10 \log\left[1 + \frac{(n-1)(\Delta P - \Delta P')}{2P_1}\right] db$$

and since

$$10 \log \frac{\Delta P}{\Delta P'} = X, \text{ then } \Delta P' = (\Delta P)\, 10^{-\frac{X}{10}} \qquad (11)$$

making this substitution, we conclude that the error in db:

$$X_m - X = 10 \log\left[1 + \frac{(n-1)\Delta P}{2P_1}\left(1 - 10^{-\frac{X}{10}}\right)\right] \qquad (12)$$

X is the attenuation in db.

Figure 2 shows the theoretical error in insertion loss measurement for various conditions of peak power, exponent n, DC bias power and decibels to be measured.

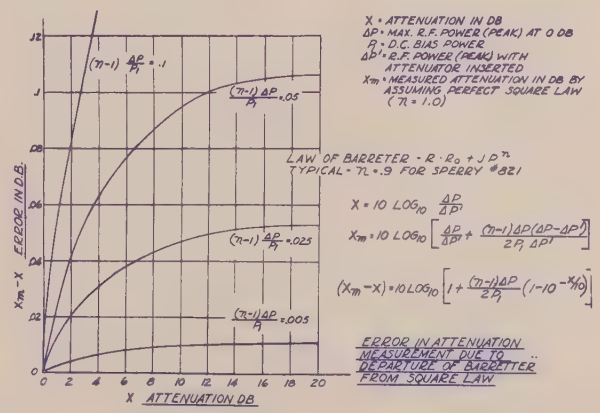

X = ATTENUATION IN DB
ΔP = MAX. R.F. POWER (PEAK) AT 0 DB
P1 = D.C. BIAS POWER
ΔP' = R.F. POWER (PEAK) WITH ATTENUATOR INSERTED
Xm = MEASURED ATTENUATION IN DB BY ASSUMING PERFECT SQUARE LAW (n = 1.0)
LAW OF BARRETER - R = R0 + JP^n
TYPICAL - n = .9 FOR SPERRY #821
X = 10 LOG10 ΔP/ΔP'
Xm = 10 LOG10 [ΔP/ΔP' + (n-1)ΔP(ΔP-ΔP')/2P1 ΔP']
(Xm - X) = 10 LOG10 [1 + (n-1)ΔP/2P1 (1 - 10^-X/10)]
(n-1) ΔP/P1 = .1
(n-1) ΔP/P1 = .05
(n-1) ΔP/P1 = .025
(n-1) ΔP/P1 = .005
Xm - X ERROR IN D.B.
X ATTENUATION DB
ERROR IN ATTENUATION MEASUREMENT DUE TO DEPARTURE OF BARRETTER FROM SQUARE LAW

Figure 2

It is self-evident that the error is a function of the attenuation to be tested. In the limit case of 0 db, a zero error will arise irrespective of how badly a bolometer is used.

In a typical case, using a P.R.D. #610-A barretter at a D.C. bias P_1 of

4 milliwatts, we limit our peak R.F. power ΔP to 200 microwatts. Since $n = .9$, $(n-1)\ \Delta P/P_1 = .005$. Consulting Figure 2, we find a maximum error of .011 db.

Experimental Derivation of Error Due To Deviation From Square Law

a) Static Characteristic

Figure 3 shows a typical D.C. character'stic of a P.R.D. 610-A barretter. This curve w s . ained using a D.C. potentiometer accurate to .01% and a standard resistor accurate to .01%. The data are therefore accurate to about .02%.

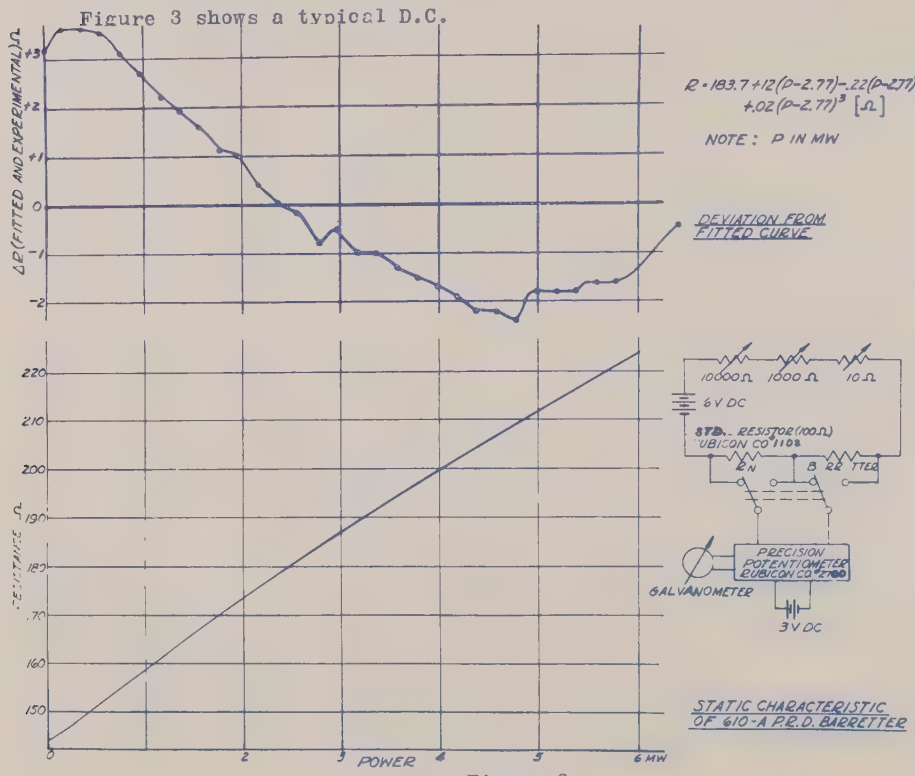


Figure 3

To obtain the deviation of the slope of this curve by visual means can lead to serious errors.

We, therefore, fitted a power series to the curve using the method of least squares.

The deviation of the fitted curve and the measured characteristic is plotted in Figure 3.

The deviation from square law in % is:

$$\delta = 100\ \frac{b_2 + 3b_3P' + b_3}{b_1}\ \frac{\%}{mW} \qquad (13)$$

Using the data obtained from our "best fitted" curve, we obtain

$$\delta = 1.5\ \frac{\%}{mW}$$

If we consider again the exponential representation

$$R = R_o + JP^n$$

the mathematical relationship between the percentage non-linearity and the exponent

n is

$$\delta = 100\left[\frac{n-1}{2P_1} + \frac{(n-1)(n-2)}{6P_1^2}\right]\frac{\%}{mW}$$

$$= 1.25\ \frac{\%}{mW}$$

If we insert n = .9, δ becomes 1.25% which is in good agreement with the above results.

b) R.F. Measurement

Figure 4 shows the block diagram of the test set-up used in the experimental verification, in order to find differences in the order of .01 db, a two-channel set-up was used. Its sensitivity and short-time stability permits estimating variations of .001 db. The reference barretter is operated at a level 10 db below the barretter under test.

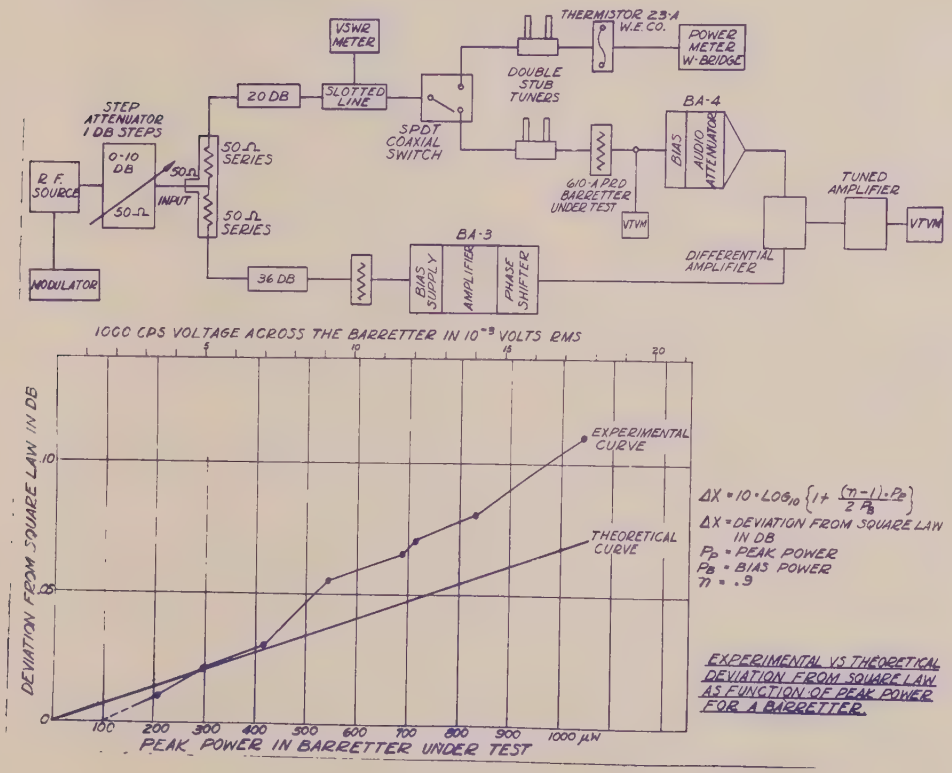


Figure 4

Figure 4 shows good agreement with the theoretically predicted deviation.

Insertion Loss Test-Sets Using Barretters

Bolometers can be used for rather accurate measurements under proper conditions.

An application of the bolometer to insertion loss measurement is shown in Figure 5. **This set produces accuracies of .1 db for a 30 db measurement.**

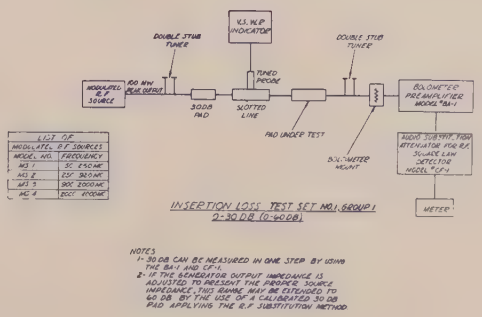


Figure 5

In order to obtain the accuracy possible, the previously listed precautions must be observed. However, there are many additional sources of error, such as, e.g.,

1. Change of modulation percentage or modulation frequency.
2. Change of R.F. level.
3. Change of R.F. frequency.
4. Change of bolometer input impedance (R.F.)
5. Bias current noise.
6. Change of transducer efficiency.
7. Variation of bolometer resistance.
8. Noise figure and bandwidth of amplifier in relation to bandwidth of indicator.
9. Non-linearity of amplifier.
10. Gain stability of amplifier.
11. Linearity of detector after amplifier (2nd detector).
12. Change of 2nd detector in presence of noise.

Conclusions

Insertion loss test sets depending upon square law detectors will find a wider application. Many circuit precautions must be designed into the equipment in order to approach the accuracies which can be realized with square law detectors.

However, the resulting maintenance and skill required to operate such test-sets compares favorably with the corresponding requirements of heterodyne-type insertion loss test-sets. For measurement of high losses such as, e.g., 80 db it will still be necessary to use heterodyne-type insertion loss test-sets. By using a 20 db or 30 db attenuator as a gage block, it is possible to extend the range of the low R.F. power square law type test set using R.F. sources providing a higher power output.

References

1. Terman, F. E., "Linear Detection of Heterodyne Signals", Electronics, 3 (1930), pp. 386-387.

2. Strutt, D., "Diode Frequency Changers", Wireless Engineer 13 (1936), pp. 73-80.

3. Gainsborough, G. F., "The Diode as Frequency-Changer for Measurements at U.H.F.", Nature, Sept. 23, 1949, pp. 548-549.

4. Colebrook, F. M. and G. H. Aston, "Diode as a Frequency Changer", Wireless Engineer, January 1943, pp. 5-14.

5. James, E. G. and J. E. Houldin, "Diode Frequency Changers", Wireless Engineer, January 1943, pp. 15-27.

6. Montgomery, C. G., "Ballantine Voltmeter Method", Technique of Microwave Measurements, 1st Ed., Vol. 11 of Radiation Lab. Series, pp. 806-808 and 838-841.

7. Silver, S.,"Microwave Antenna Theory and Design", 1st Ed., Vol. 12 of Radiation Lab.Series, pp. 599, 602-606.

8. Korewick, John, "Audio Modulation Substitution System for Microwave Attenuation Measurements", TRANSACTIONS OF THE I.R.E., Pg. MTT, Vol. MTT-1, No. 1, March 1953, pp. 14-21.

ENGINEERING NOTES NO. 3
April 1955

THE THERMAL TIME CONSTANT OF A BOLOMETER

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This paper was presented during the Radio Conference on September 8, 1954 at the Boulder Laboratories of the National Bureau of Standards.

Introduction

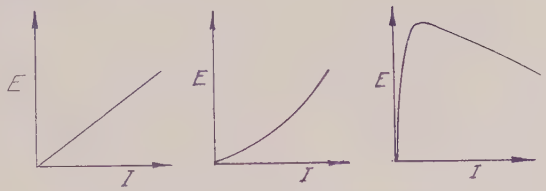
A convenient method for measuring R-F power ratios with audio frequency techniques is the following:[1,2,3]

R-F power is modulated with audio frequency and dissipated in a bolometer. The resistance of the bolometer changes then with the audio frequency and by feeding the bolometer with a constant DC current an AC voltage across the bolometer is produced which is amplified and detected. The bolometer must be able to change its resistance sufficiently fast so it can follow the modulated power. This ability is limited by the thermal time constant of the bolometer.

The first part of this paper deals with the theory of the time constant of bolometers,[4,5] the second part discusses two methods of measurement and the third part treats a theory of multiple time constants including their distribution.

The Theory of the Thermal Time Constant

Plotting voltage versus the current for a usual resistor produces a straight line, if the resistor is physically large enough so that its power dissipation does not change its temperature appreciably. If the resistor is very small (e.g., the more familiar barretter is a thin platinum wire about 30 microinches diameter, the usual R.F. thermistor is a small semi-conductor bead about 15 mils diameter), the temperature increases noticeably with the power dissipated. The change in temperature will affect the resistance and we obtain the typical thermistor and barretter curves as shown in the D.C. characteristics of Figure 1.



Large Resistor | Barretter | Thermistor

D. C. Characteristics
Figure 1

The increase in temperature occurs in a finite time, determined by the power dissipated, heat capacity and rate of heat loss of the bolometer. The smaller the heat capacity of the bolometer, the shorter is the time.

For derivation of the differential equation relating the resistance of the bolometer, the time, and the input power one considers first the heat balance equation:

$$P = P_o + C_p \cdot \frac{\partial T}{\partial t} \qquad (1)$$

P = electrical power input in watts
P_o = thermal heat loss, expressed in watts
C_p = heat capacity of the bolometer in Joules/degree°C
T = temperature in°C
t = time in seconds

Expressing R as a power series in T while neglecting higher order terms results in equation (2).

$$R = R_o + \alpha (T - T_o) \qquad (2)$$

Since only T is a function of time, one obtains

$$\frac{\partial R}{\partial t} = \alpha \cdot \frac{\partial T}{\partial t} \qquad (3)$$

and

$$\frac{\partial T}{\partial t} = \frac{1}{\alpha} \cdot \frac{\partial R}{\partial t} \qquad (4)$$

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Considering a small temperature increase and, therefore, neglecting radiation losses and the deviation of the convection loss from **linearity**, one may state:

$$P = K(T - T_0) \qquad (5)$$

where K is a measure of the heat conductivity of the bolometer.

Then:

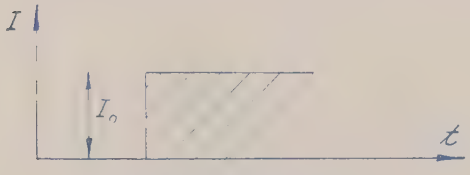
$$P_0 = \frac{K}{\alpha}(R - R_0) \qquad (6)$$

$$P + \frac{K}{\alpha} R_0 = \frac{K}{\alpha} R + \frac{C_p}{\alpha} \frac{dR}{dt} \qquad (7)$$

This is the desired differential equation, relating the resistance of the bolometer, the time, and the power input, which must be solved for various conditions of power input as a function of time.

2. Measurement of Thermal Time Constant by Step Current Function for Bolometers with Large Time Constants.

For a bolometer of large time constant, it is convenient to determine the time constant directly by applying a current as a function of time of this shape:



Current Step Function
Figure 2

The boundary conditions applicable to equation (7) are

$$t = 0;\ R = R_0 \text{ AND } t = \infty;\ \frac{dR}{dt} = 0 \qquad (8)$$

Replacing P in equation (7) by I^2R, one obtains:

$$\frac{K}{\alpha} R_0 = \left(\frac{K}{\alpha} - I^2\right) R - \frac{C_p}{\alpha} \frac{dR}{dt} \qquad (9)$$

The solution of equation (9) is:

$$R = \frac{R_0}{1 - I^2 \frac{\alpha}{K}} + \frac{R_0}{1 - \frac{K}{I^2 \alpha}} e^{-\frac{t}{\tau}} \qquad (10)$$

with the time constant:

$$\tau = \frac{C_p}{K - I^2 \alpha} \qquad (11)$$

This reveals that the time constant also depends upon the current. Multiplying R by the current I_0, produces the voltage across the bolometer, which can be observed on an oscillograph. From this measurement, one can determine the time constant for the constant direct current I_0.

3. Time Constant Measurement by Sine Waves for Bolometers with Small Time Constants.

A combination of direct current I_0 and an audio frequency current, is represented as

$$I = I_0 + a \cdot e^{j\omega t} \qquad (12)$$

This may be applied to equation (7). In order to obtain the steady state solution, one may assume:

$$Z = \bar{R} + Z_V e^{j\omega t} \qquad (13)$$

R was changed to Z since Z is complex. This produces

$$\frac{K}{\alpha} R_0 = \left(\frac{K}{\alpha} - I_0^2 - 2 I_0 a e^{j\omega t}\right)\left(\bar{R} + Z_V e^{j\omega t}\right) + \qquad (14)$$
$$+ \frac{C_p}{\alpha} j\omega Z_V e^{j\omega t}$$

The solution is:

$$Z = \bar{R} + R_d \frac{1}{1 + j\omega\tau} \qquad (15)$$

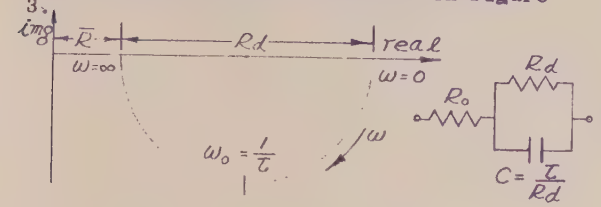
where:

$$\tau = \frac{C_p}{K - I_0^2 \cdot \alpha} \text{ AND } R_d = \frac{2 I_0 \cdot a \cdot \bar{R}^2}{R_0} \frac{\alpha}{K} \qquad (16)$$

In the particular case of barretters, one may write; because α is positive:

$$R_d = \frac{2 I_0 a \bar{R}^2}{R_0} \frac{|\alpha|}{K} \qquad (17)$$

The impedance diagram and equivalent circuit for a barretter are shown in Figure 3.



Barretter Impedance Diagram
And Equivalent Circuit
Figure 3

In the case of the thermistor, one may write

$$R_d = -\frac{2 I_0 \cdot a \bar{R}^2}{R_0} \cdot \frac{|\alpha|}{K} \qquad (18)$$

because α is negative.

We obtain

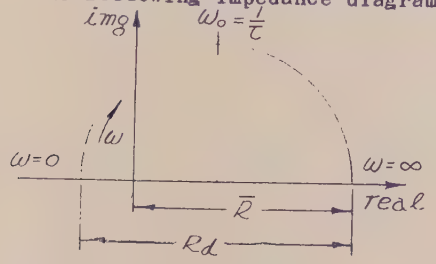
$$Z = \bar{R} - |R_d| + |R_d| \frac{j\omega L}{|R_d| + j\omega L} \qquad (19)$$

and

$$\tau = \frac{L}{|R_d|} \qquad (20)$$

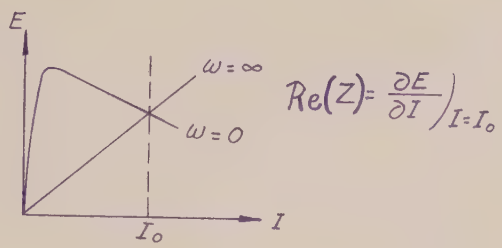
Because $|R_d|$ is larger than $\bar{R}$,

$\bar{R} - |R_d|$ is negative and we obtain the following impedance diagram:



Impedance Diagram of Thermistor
Figure 4

The impedance for $\omega = 0$ and $\omega = \infty$ can also be obtained from the corresponding E vs. I curves



Thermistor Characteristic
Figure 5

These equations permit the measurement of response of bolometers to different time functions of current or power. These measurements can be used to forecast the behaviour of the bolometer to modulated RF power by using simple and common AC bridge methods at low frequencies. The measurements reveal not only the real part of the frequency dependent part of the bolometer impedance, but also the imaginary part which is necessary for a proper determination of the time constant. We know this imaginary part has the following frequency function and has a maximum value at $\omega_0 = \frac{1}{\tau}$

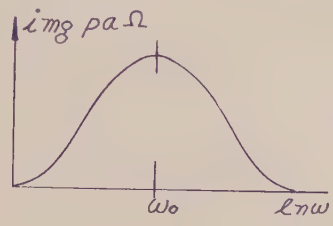
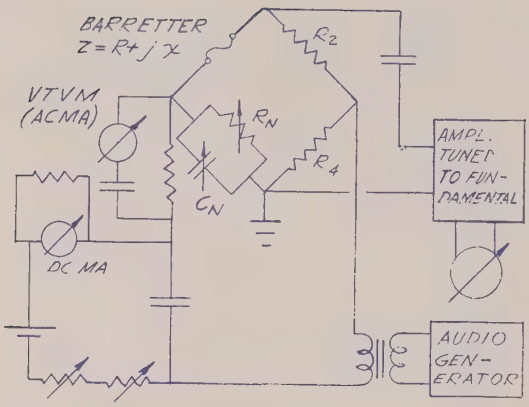


Figure 6

4. Measurements.

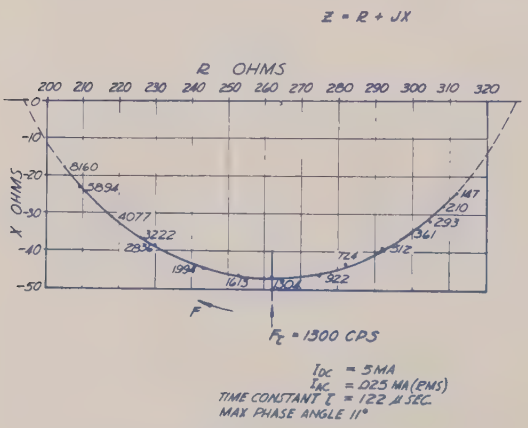
The impedance of a barretter as a function of frequency can be measured by a circuit, such as that shown in Figure 7.



A. F. Bridge
Figure 7

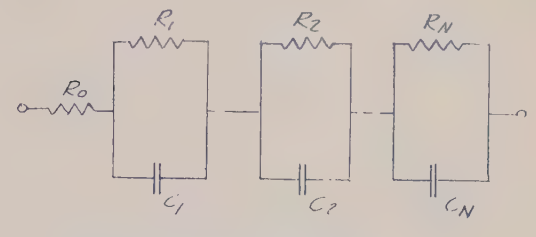
For thermistor measurements, one may use a Maxwell bridge, since the reactive component of the thermistor impedance is positive. That permits putting the RC-combination in the other bridge branch across the thermistor.

Figure 8 shows a typical curve measured for a PRD 610-A barretter.



Impedance Diagram of A
Barretter P.R.D. 610-A
Figure 8

The impedance diagram is not the expected semicircle, but resembles an ellipse. Such curves appear also in the study of dielectric and magnetic relaxation effects,[6,7,8,9,10] where the simple theory is mathematically the same as prescribed here and predicts a semicircle. In the case of dielectric and magnetic relaxation, the ellipticity of the curve is explained by the fact that there is not only one, but many time constants. It is very reasonable to assume that the same effect is present here; that is, that the equivalent circuit is more accurately given by Figure 9.



Equivalent Circuit For
Multiple Time Constant.
Figure 9

Theory of Time Constant Distribution

Divide a bolometer wire physically in small elementary domains and give each domain a certain time constant. Then one can say that each domain contributes a part dZ to the whole impedance Z :

$$dZ = R_d \cdot g(\tau) \cdot d(\ln \tau) \cdot \frac{1}{1+j\omega\tau} \qquad (21)$$

$g(\tau)$ is the distribution function of the time constant for all the domains.

To obtain the impedance Z, one must integrate over all time constants:

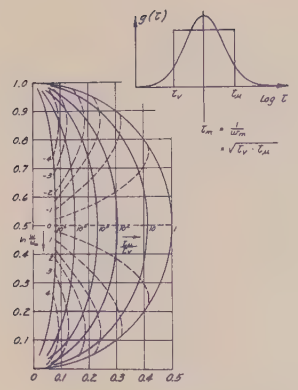
$$Z = R_d \int_0^\infty g(\tau) \frac{d(\ln \tau)}{1+j\omega\tau} \qquad (22)$$

This complex integral can be solved for certain time distribution functions in a closed form.

Figure 10 shows the computed impedance diagram for a Gaussian distribution:

$$g(\tau) = \frac{b}{\sqrt{\pi}} \cdot e^{-b^2(\ln \tau)^2} d(\ln \tau) \qquad (23)$$

The number b represents the dispersion of τ. To make this dispersion a little more evident, one approaches the distribution curve by the best fitting rectangular curve and takes the ratio of the largest to the smallest time constant as dispersion.



Computed Impedance Diagram For Gaussian Distribution of Time Constants. Figure 10

Comparing the measured curve shown in Figure 8 with the computed curves, one finds a distribution of

$$\frac{\tau_\mu}{\tau_v} \approx 50 \qquad (24)$$

Change of The Time Constant With The Current

Equation (11) reveals that the time constant changes with the current flowing through the bolometer. One can estimate how large this influence is for a barretter and a thermistor.

Considering the 610-A (PRD) barretter, the following constants apply:

$$\frac{\alpha}{K} = 12\,\frac{\Omega}{mW}\ ;\ I_o = 4\,ma;\ I_o^2\frac{\alpha}{K} = .19 \qquad (25)$$

This indicates that the influence is small. Changing the D.C. bias from 0 to 4 Ma, produces a change of the time constant of only 20%.

Considering a Western Electric 23-A thermistor, α is negative.

Therefore:

$$\tau = \frac{C_p}{K}\cdot\frac{1}{1+I_o^2\frac{|\alpha|}{K}} \qquad (26)$$

The following constants apply:

$$\frac{|\alpha|}{K} = 40\,\frac{\Omega}{mW};\ I_o = 6\ ma;\ I_o^2\frac{|\alpha|}{K} = 1.44 \qquad (27)$$

One sees here that the time constant of a thermistor is very closely connected to the current flowing through it.

Response of the Bolometer To Modulated R.F.

A very useful application of bolometers is detecting the audio modulation of an R.F. source. The bolometer is carrying a D.C. current and is dissipating a modulated R.F. power. That means one must insert a power in this differential equation:

$$P = I_o^2R + P_R\left(1+e^{j\omega t}\right) \qquad (28)$$

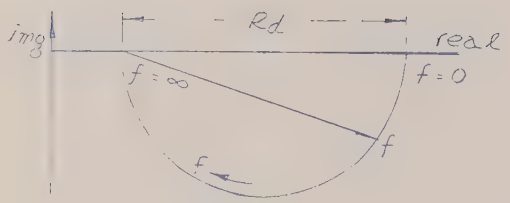
Solving this equation for Z_V, one obtains:

$$Z_V = \frac{P_R}{\frac{K}{\alpha}-I_o^2}\cdot\frac{1}{1+j\omega\,\frac{C_p}{K-I_o^2\alpha}} \qquad (29)$$

That means that the frequency dependance of Z_V will be the same for a sinusoidal R.F. power impressed on the bolometer as that of a sinusoidal current superimposed on the D.C. current. Then the voltage across the bolometer will be:

$$E = I_o\,|Z_V| = \frac{P_R\,I_o}{\frac{K}{\alpha}-I_o^2}\left|\frac{1}{1+\omega^2\tau^2}\right|^{\frac{1}{2}} \qquad (30)$$

Since the magnitude of Z_V is the only frequency dependent part of this equation, one may conclude that the dependence of the output of the bolometer on frequency is the same as the dependence upon frequency of the magnitude of the variable part of the complex impedance, when this impedance is measured with an A.C. bridge.



Frequency Dependence of
Barretter Impedance
Figure 11

Referring to Figure 11, one may define as frequency response the ratio

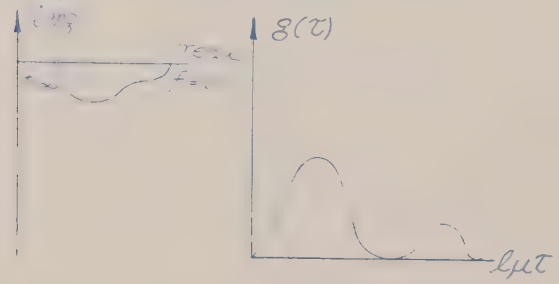
$$\frac{\text{output voltage at } f=0}{\text{output voltage at } f=f} \qquad (31)$$

or in db:

$$\text{response} = 20 \log \frac{\infty - f}{Rd} \qquad (32)$$

This permits computation of the frequency response from the impedance diagram. One also sees that it is nearly impossible to obtain the true time constant by measuring only the voltage across the bolometer, when we feed a modulated RF power in and change the modulation frequency, in spite of the difficulties of this type of measurement. For the ideal case of only one time constant, this would be easier, but usually one finds a certain time constant distribution, which even sometimes may be non-uniform consisting of several main time constants.

Figure 12 shows a typical experimentally determined diagram and the corresponding time constant distribution.

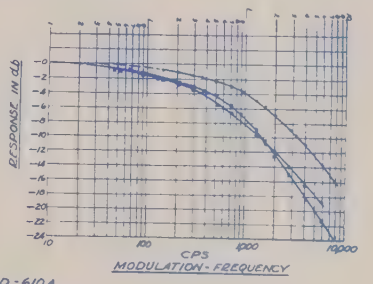


Typical Barretter Impedance Diagram — Corresponding Time Constant Distribution

Figure 12

The 3 db point of the frequency response, which is usually used for the determination of the time constant, is no longer the right point. For the 610A barretter, shown in Figure 8, the correct point is at 4.8 db. The impedance diagram, however, gives a complete picture of the behaviour of a bolometer. This is also useful for the design of bolometer bridges. Bolometer-bridges are mostly used at audio-frequencies, because it is easier to detect small ac signals than small dc signals. Knowing the impedance diagram, one knows the complex impedance of the bolometer at the frequency of interest and can therefore calculate the other bridge branches.

Figure 13 shows the frequency response calculated from the impedance diagram, for a PRD-610A, a Sperry 821, and a Narda 821.



BARRETTER EFFICIENCY
VS
FREQUENCY

Figure 13

Figure 14 shows the response of a barretter to a squarewave modulated R.F. power at different modulation frequencies as well as the time function of the RF power, obtained by means of a 1 N 21 crystal.

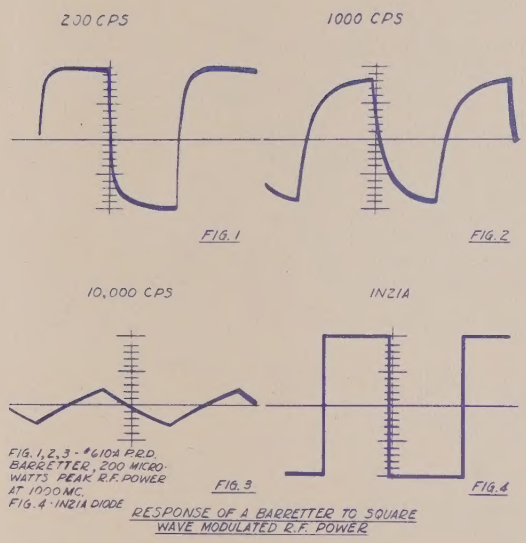

200 CPS
1000 CPS
FIG. 1
FIG. 2
10,000 CPS
IN21A
FIG. 1, 2, 3 - #610A P.R.D. BARRETTER, 200 MICRO-WATTS PEAK R.F. POWER AT 1000 MC.
FIG. 4 - IN21A DIODE
FIG. 3
FIG. 4
RESPONSE OF A BARRETTER TO SQUARE WAVE MODULATED R.F. POWER

Figure 14

Acknowledgements

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References

1. "Ballantine Voltmeter Method", pp. 806-808 and pp. 838-841, Vol. 11 of Radiation Lab. Series (Technique of Microwave Measurements, by C. G. Montgomery, 1st Edition).

2. pp. 599, 602-606, Vol. 12 of Radiation Lab. Series (Microwave Antenna Theory and Design, by S. Silver, 1st Edition).

3. "Audio Modulation Substitution System for Microwave Attenuation Measurements", by John Korewick, pp. 14-21, March 1953, Transactions of the I.R.E., Pg. MTT, Vol. MTT-1, No. 1.

4. "Validity of the Substitution Principle in Power Measurement" by Max Sucher and Herbert J. Carlin, September 24, 1951, Report 257-51, PIB-198, Contract No. AF 30(120)-490.

5. "The Theory of Operation of a Barretter", pp. 161-168, Vol. 11 of Radiation Lab. Series, (Technique of Microwave Measurements, by C. G. Montgomery, 1st Edition).

6. "Annalen der Physik", by K. W. Wagner, Vol. 40, 1931, p. 817.

7. "Frequenz" by H. Wilde, Vol. 3, 1949, pp. 309 and 348.

8. "Zeitschrift f. Angew Physik", by R. Feldtkeller and H. Hettich, Vol. 2, 1950, p. 494.

9. "Zeitschrift f. Angew Physik", by Th. Einsele and F. Baur, Vol. 3, 1951, p. 373.

10. "Archiv Fuer Electrische Uebertragung," by R. Feldtkeller and G. Sorger, Vol. 7, 1953, p. 79.

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